ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

FEDERAL TELEPHONE AND RADIO CORPORATION: 1919–1946

COLOR-TELEVISION TRANSMITTER FOR 491 MEGACYCLES

SQUARE LOOPS FOR FREQUENCY-MODULATED BROADCASTING

TRIODE AMPLIFICATION FACTORS

ATTENUATION AND Q FACTORS IN WAVE GUIDES

PIEZOELECTRIC SUBSTANCES

ELECTRON TRAJECTORIES IN A PLANE SINGLE-ANODE MAGNETRON

SELENIUM RECTIFIERS FOR BROADCAST RADIO RECEIVERS

SINGLE-UNIT RADIO EQUIPMENT FOR PASSENGER AND CARGO VESSELS

MEASURING GRID PRIMARY EMISSION IN THERMIONIC VALVES

RECENT TELECOMMUNICATION DEVELOPMENTS

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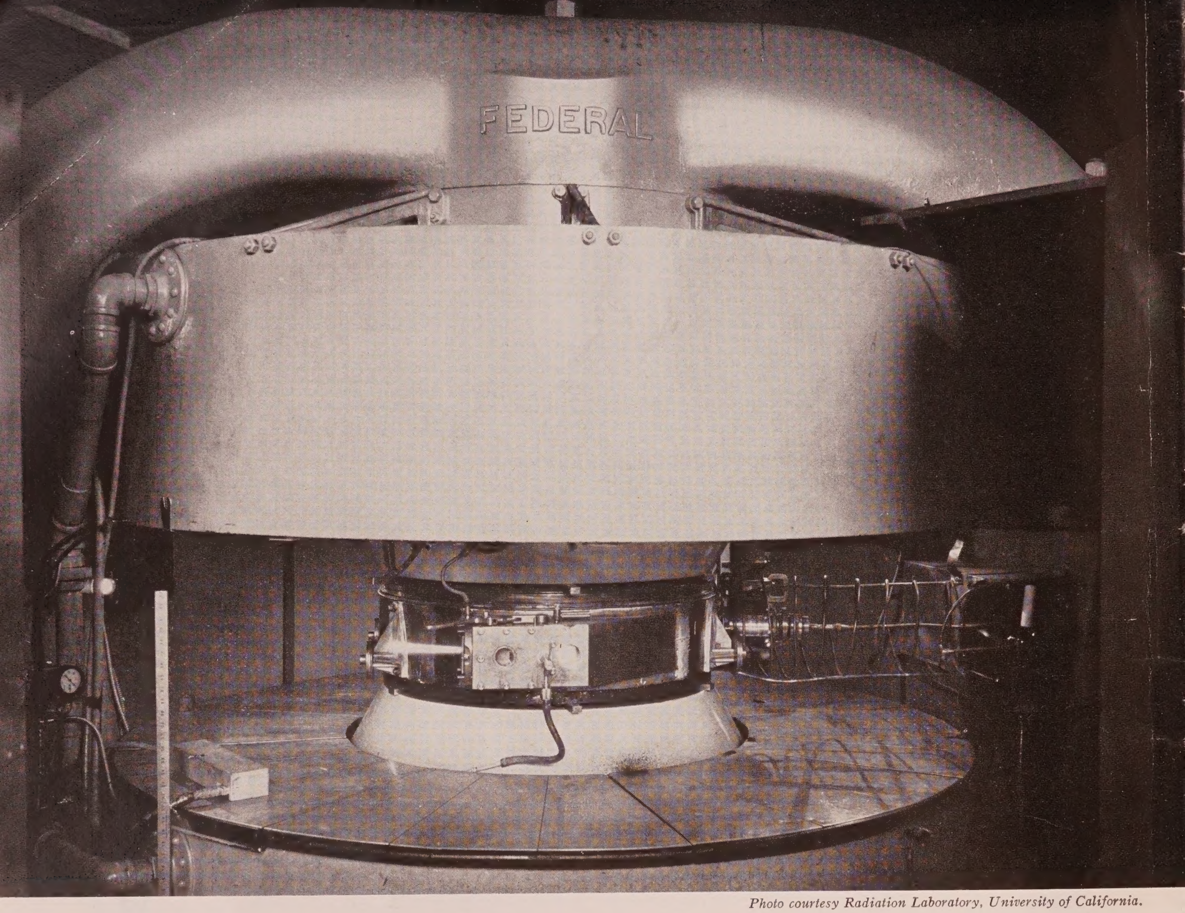
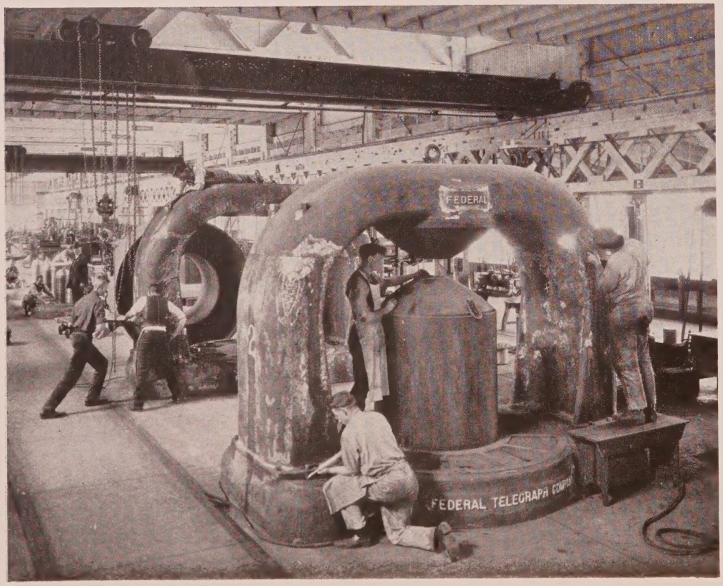


Photo courtesy Radiation Laboratory, University of California.



The illustration above shows the first giant cyclotron ever built. It was constructed under Dr. E. O. Lawrence's direction at the University of California. When completed, it weighed 80 tons and employed a 65-ton magnet cast originally for a Federal 1000-kilowatt arc converter. At the left is shown a view of work being completed on two of the magnet castings in the Federal factory at Palo Alto, California, in 1921. The magnet fields were originally cast in 1918 for a World War I project stopped by the Armistice. They were later intended for a radio station in China. Japanese opposition prevented completion of the Chinese station.

Dr. Leonard Fuller, while vice president of Federal, learned of Dr. Lawrence's need of a magnet for producing a practical cyclotron. Federal, consequently, donated the casting to the University and also assisted in winding the magnet coils.

Thus, to Federal's long history of scientific pioneering, can be added this contribution toward the disclosure of the secrets of atomic power.

Federal Telephone and Radio Corporation

A Historical Review: 1909–1946

By F. J. MANN

Managing Editor, Electrical Communication

FEDERAL had its beginning in much the same manner as so many great enterprises. First there was someone with an idea. When that idea became a reality, it served as the means to a growth which dwarfed even the imagined scope of the original concept.

The man responsible for the founding of Federal was C. F. Elwell, the company's first director and chief engineer. Elwell's idea was to find a practical method of transmitting voice by radio. His search led him far afield, but it resulted in founding an organization that later became responsible for the discovery and development of a number of outstanding electronic and radio devices.

In 1908, Elwell was a tall, athletic young man completing the course in electrical engineering at Stanford University, Palo Alto, California. He had been collaborating with Professors G. H. Clevenger and D. A. Lyon on the design of an electrical furnace for the reduction of iron on a commercial scale. This work was selected by Elwell for his graduation thesis and he was devoting all of his time to it. He had not then even thought of going into radio work although he has spent his entire life since in the field of radio engineering.

The McCarty Wireless Telephone

Elwell was just about to make a trip east to purchase three large transformers for the new furnace when he was approached to do research on the McCarty wireless telephone patent which had been purchased by the Henshaw brothers, bankers of Oakland, California. To persons familiar today with high-fidelity radio broadcasting from transmitters ranging in powers up to 200 kilowatts, the idea that anyone would consider the McCarty patent seriously for radiotelephone transmission seems preposterous. For this patent covered a device intended to permit radiotelephone transmission by means of a spark coil which was to be equipped with two primaries, one for a microphone and the other for the make and break. This spark coil was to be part of a conventional, damped-wave spark transmitting circuit employing a fixed spark gap, capacitor, inductor, and antenna and ground.

When a position to experiment with the McCarty patent was first offered Elwell, he turned it down because of the pressure of work on the electric furnace. He might also have rejected the proposal because of the hopelessness of expecting such a crude piece of apparatus to operate. However, no one at that time had proved whether it would or would not work and at least several persons including the U.S. Patent Office thought it might. Elwell was invited a second time by the representative of the Henshaw brothers to undertake the research. The Henshaw brothers were anxious to learn whether the patent, on which they had already spent a considerable sum of money, was of practical value. This second offer was sufficiently attractive to cause him to agree to investigate the possibilities of the spark-coil radiotelephone transmitter.

While on his trip east for the furnace transformers, Elwell purchased the necessary apparatus for the McCarty transmitter. He returned to Palo Alto in the summer of 1908 and began work on the McCarty patent at Stanford University. Since Mr. Timothy Hopkins, a trustee of the university who was in charge during the summer, objected to these experiments being carried on there, Elwell had to leave. He therefore arranged for the purchase of a house in Palo Alto located on the corner of Cowper Street and Embarcadero Road.

Elwell erected two 75-foot wooden masts to support his transmitting aerial on the new property. The house he used as a laboratory and dwelling. The receiving station was located in a small wooden hut placed near the base of one of the nearby Palo Alto water towers which served to support the receiving antenna.

It Worked—but Not Well

Two-way transmission and reception was not attempted since it was possible to see the receiving hut from the transmitting station with the aid of a pair of binoculars. Elwell reports that he actually did succeed in transmitting telephone signals using this crude apparatus. As a matter of fact, the incident is recalled vividly to mind because, as Elwell states, he had arranged with his assistant at the receiving station to signal the results of the experiments with a piece of paper placed at one of the four corners of the hut. The position of the paper was intended to indicate the degree of reception. The assistant, however, used such a small piece of paper that Elwell could not see it through the binoculars. When Elwell returned to the microphone he shouted impatiently, "Why don't you put up a piece of paper I can see?" This time the assistant came out with a whole sheet of newspaper!

Demonstrations given on August 29, 1908, were reported in the San Francisco, Oakland, and Palo Alto newspapers in glowing terms. Like flying and other new inventions of the time, the idea of telephoning without wires had appealed to the popular imagination and so any success at all was good newspaper copy. But Elwell had to report to his employers that the McCarty patent was not practical for commercial exploitation.

This work on the transmission of voice with damped waves had completely diverted Elwell's interest from the electric furnace to radio. He had noticed that the few times he did obtain results with the McCarty transmitter occurred when he placed the spark gaps so close together that a small arc discharge took place, which gave slightly damped or nearly continuous waves. With the gaps so arranged it was possible to transmit intelligible speech and music over distances up to five miles. Elwell became convinced that wireless telephony would only be commercially practical when a reliable system of generating undamped or continuous waves of sufficiently high frequencies could be developed.

Early Sources of Continuous Waves

Although Dr. Lee de Forest had invented his audion in 1906, it was not until later when he was working in the Federal Palo Alto laboratories that he devised circuits for the oscillating audion. At the time, therefore, only two sources of undamped or continuous waves were known—the high-frequency alternator, then only in the development stage, and the Poulsen arc. Several one-kilowatt alternators had been made in the U. S. A. by Alexanderson to operate at a frequency of 50000 cycles. It was possible to transmit voice with these transmitters, but the extremely long wavelength made them impractical except for rather elaborate and costly installations.

Valdemar Poulsen, the Danish scientist, invented the arc that bore his name in 1903. The original work on the arc was done in 1892 by Elihu Thomson who showed that direct current could be employed for the generation of oscillatory currents if capacitance and inductance were placed across a spark gap connected to the supply mains through large choke coils and a magnetic blast provided for the spark gap. Thomson obtained small currents at frequencies up to 100000 cycles. In 1900, William Duddell substituted carbon electrodes for the spark gap and obtained continuous oscillations at audible frequencies. By immersing this arc in hydrogen gas, Poulsen was able to produce higher frequencies and to generate more power. The frequencies thus produced were practical for the transmission of radio signals.

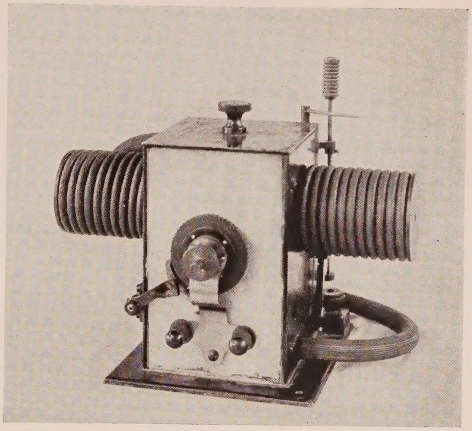
Some reports of Poulsen's work had reached the United States by 1908. While in these papers, published in scientific and semiscientific journals (some of translations from speeches made by Poulsen), there were indications of the potentialities of the Poulsen continuous-wave generator, very little had been done to make use of the invention commercially. A syndicate was formed in Copenhagen to exploit the Poulsen arc patents and develop them for radiotelegraphy and radiotelephony. A station was erected at Lyngby, Denmark, and later a second station was set up at Esbjerg, 180 miles away. Over these stations, it had been possible to carry on radio communication, but little more had been done with the Poulsen arc by 1909.

Predominance of the Spark Transmitter

The vast expanse of the North American continent seemed to offer great inducements to long-distance radio development in the U. S. A. No one had even thought of radio broadcasting, but the search for a practical method of radiotelephone transmission was inspired by the idea then prevalent that the radio someday would replace wire telephony. Even Fessenden, an outstanding radio engineer and inventor of the time, wrote in 1909 in the *Journal of Electricity, Power and Gas* that radiotelephony would no doubt replace wire telephony, at least between cities. And yet, of 475 radio stations then in operation in the United States (1909), all utilized the noisy spark transmitter.[1] As the size of these transmitters grew, the spark-noise became louder and louder so that radio rooms on ships were built like iceboxes to keep the sound from disturbing passengers. Advocates of the spark transmitter were almost fanatical in its praise.

When Elwell reported to the Henshaw brothers that the McCarty patent was worthless, he had already studied all the available literature on Poulsen's invention, which at that time was scant. Elwell had even cabled Poulsen to ask if he had sold his U. S. patent rights and, if not, what would be his terms. Poulsen cabled back promptly mentioning a price of a quarter of a million dollars and stating other conditions. But when Elwell suggested working with the Poulsen arc to the Henshaw brothers they turned the suggestion down, saying they had had enough of wireless.

The reluctance of the Henshaw brothers to invest further in the wireless business at the time is understandable. First, probably as much money was being sunk in worthless radio patents as in never-to-be-productive oil wells. The whole idea of communication without wires had so captivated the human mind that radio was an easy source of stock exploitation and speculation. Even at that early date, when the art was so crudely developed, money was being accepted for stock in television schemes. Further, the more solid citizenry placed great reliance on the spark transmitter and its possibilities. These were the persons who were still driving sleek horses and buggies and laughing at the noisy, ugly automobile. Science had not yet reached the stage of precision where the new development could be evaluated in advance. So, like the horse and buggy, the spark transmitter was regarded as doing a good and practical job, noisy as it was, and the Poulsen arc was considered strange and just a little unbelievable.


First Poulsen arc brought to the U.S.A. from Denmark in 1909. Although rated at 100 watts, it actually delivered about fifty watts of radio-frequency power. With it, radiotelephone signals were sent from North Palo Alto to South Palo Alto, thereby demonstrating the efficacy of the Poulsen arc for radiotelephony. The early demonstrations with this arc led to the formation of Poulsen Wireless Telephone and Telegraph Company, predecessor of Federal.

Elwell Goes to Denmark

In view of all this skepticism, Elwell felt he should go to Denmark to see for himself what Poulsen could do. So in May, 1909, at his own expense since he could obtain no financial backing, Elwell travelled to Copenhagen. There he met Dr. Poulsen and Dr. P. O. Pederson,[2] Poulsen's collaborator. The two Danish scientists took Elwell to a laboratory in Copenhagen where he heard the wireless telephone signals of the Lyngby station ten miles away. Elwell reports that the articulation was perfect although the signals were weak.

The next day Elwell had his first view of the Poulsen arc at Lyngby. While at the Lyngby station, he also heard signals from the Esbjerg station 180 miles away. At the same time, Poulsen also showed him photographic recordings of dot and dash signals made at 200 words per minute. Because of this new rapid telegraph system, it was explained to Elwell, the price of the U. S. rights to the patents had been raised to $450,000.

---

[1] Robert H. Marriott, "United States Radio Development," *Proceedings of the I.R.E.*, v. 5, pp. 179–198; June, 1917.

[2] "Peder Oluf Pedersen," *Electrical Communication*, v. 20, n. 2, pp. 133–135; 1941.

Elwell was convinced that the Poulsen patents offered great possibilities for development in the U. S. A. He could see now that the silent, steady Poulsen arc, in conjunction with the high-speed telegraph system, had practical advantages for wireless telegraphy capable easily of overshadowing the noisy and slow spark station. Therefore this telegraphic development could and did become even more important than the radiotelephone Elwell had travelled to Europe to see and hear. Even with all these possibilities, the decision of a young man just out of college and with no financial backing to take the rights to the Poulsen patents back to America was, to say the least, ambitious.

Buys Rights to Poulsen Arc

Elwell negotiated with the Poulsen interests to buy the patents and, after investigating him carefully through the American consul in Copenhagen, they agreed to an option for only $1,000 in cash. Elwell also agreed to purchase two Poulsen transmitters, one rated at 5 kilowatts and one at 12 kilowatts, for a total of $6,000, half payable as soon as he returned to the U. S. A. and half when the sets were ready.

Elwell returned to New York elated over obtaining the contract and full of hope of raising money necessary to form a company to develop the system commercially. He went to Wall Street and met Thomas Fortune Ryan and other leading financiers, but by then these men were very shy of wireless schemes. There had been so much stock jobbing in damped-wave or spark wireless systems that the market in them was extremely bad. Elwell decided from this reception that the option obtained in Denmark was going to be very difficult to finance.

Unshaken in his faith in the Poulsen arc, but realizing that financial backing would not be forthcoming so easily as he had originally hoped, Elwell that summer made a second trip to Denmark. He was able to renegotiate the contract on easier terms requiring less cash and a greater amount in stock in the new company to be formed. With this new contract, he returned directly to Palo Alto.

On his second trip to Denmark, Elwell had purchased a small Poulsen arc extravagantly rated at 100 watts. It consisted chiefly of a small hollow cube formed by two sides of marble with the other two sides, top, and bottom, of brass. The anode was mounted in the center of one of the marble slabs and the cathode in the center of the other. Thus the marble acted as insulator to the electrodes. It worked on 110 volts, d.c., giving about one ampere. It was this arc Elwell took to Palo Alto, the first Poulsen arc in America.

First Arc Telephone Conversations in America

With the small Poulsen arc, Elwell constructed a wireless telephone on which he was able to transmit clear telephone signals from North Palo Alto to South Palo Alto, a distance of about one mile. This equipment convincingly proved the efficacy of continuous waves for radiotelephony and completely overshadowed the previous crude attempts with damped waves using a spark coil as a generator.

Present at these first demonstrations were David Starr Jordan, president of Stanford University, Dr. C. D. Marx, head of the civil engineering department, Prof. C. B. Wing of the civil engineering department, Prof. L. M. Hoskins, head of the mathematics department, and many other members of the Stanford University faculty as well as prominent citizens of Palo Alto. All of these men were impressed by the demonstrations and convinced of the commercial value of the arc. When Dr. Jordan offered to put up $500 toward starting a company, a number of other faculty members and Palo Alto financiers followed with financial support.

In October 1909, the Poulsen Wireless Telephone and Telegraph Company, direct predecessor of Federal, was formed as a California corporation. Assisting in the incorporation were R. W. Barrett, F. A. Wise, and M. A. Thomas, all Stanford University graduates. Among the first directors were Barrett and Professor Marx. Elwell was made director and chief engineer. The new company set up offices in the Foxcroft Building, 68 Post Street, San Francisco, California.

Although some cash had been collected in connection with the incorporation, a great deal more money was needed to purchase the patent rights

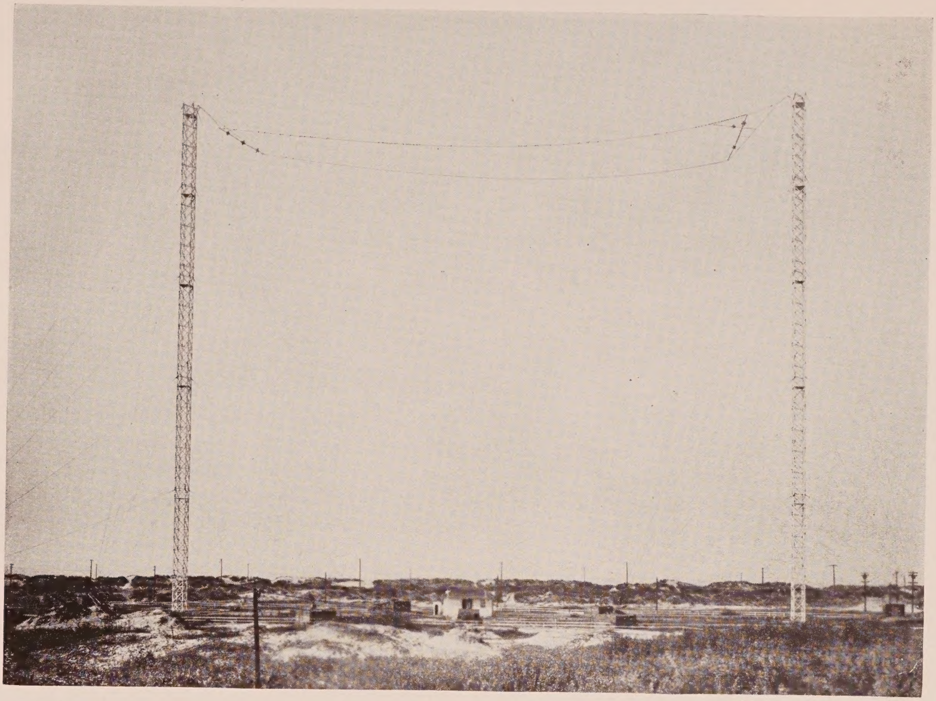


Photo courtesy C. F. Elwell

The two 300-foot wooden lattice masts of the old "Beach" station erected near Ocean Beach, San Francisco, in 1910. Lengths of cables previously used to propel the old Geary Street cable cars in San Francisco provided inexpensive guys for the two masts which were the first of their kind built in the U.S.A. The first paid commercial messages handled by the Federal Telegraph Company were sent from this station.

and to finance manufacturing facilities. Stock had been issued and it was planned to sell shares to small investors since large financial interests had refused to support this new radio venture.

Stockton and Sacramento Stations

Because so many radio stock ventures had proved failures and since it was still an age when seeing was believing, it was decided to build two stations at least fifty miles apart for radiotelephone demonstrations. Sites in Stockton and Sacramento, California, were chosen for the first two radio stations to employ the new system. It was planned to use the transmitting apparatus purchased from Denmark and which was scheduled for early delivery to the U. S. A.

Two 180-foot masts were erected at each site, each mast consisting of 90-, 75-, and 60-foot poles spliced together. These poles had to be towed to each city behind the river steamer which then traveled regularly from San Francisco, because they were too long to be taken on board.

The two Poulsen transmitters finally arrived in New York and the company, from its small financial resources, was forced to pay $4,500 duty on them before they could be shipped to Palo Alto. In addition, the fare and expenses of three Danish engineers had to be provided so that they could come to the U. S. A. and install the new apparatus. The three engineers, Peter V. Jensen, F. Albertus, and C. L. Schou, duly arrived and set up the apparatus at Sacramento and Stockton early in February, 1910. It is interesting to note that Jensen and Albertus left the company, along with E. S. Pridham, one of the company's first engineers, shortly after to go into business for themselves. The name of Jensen is now well known as a manufacturer of loudspeakers,

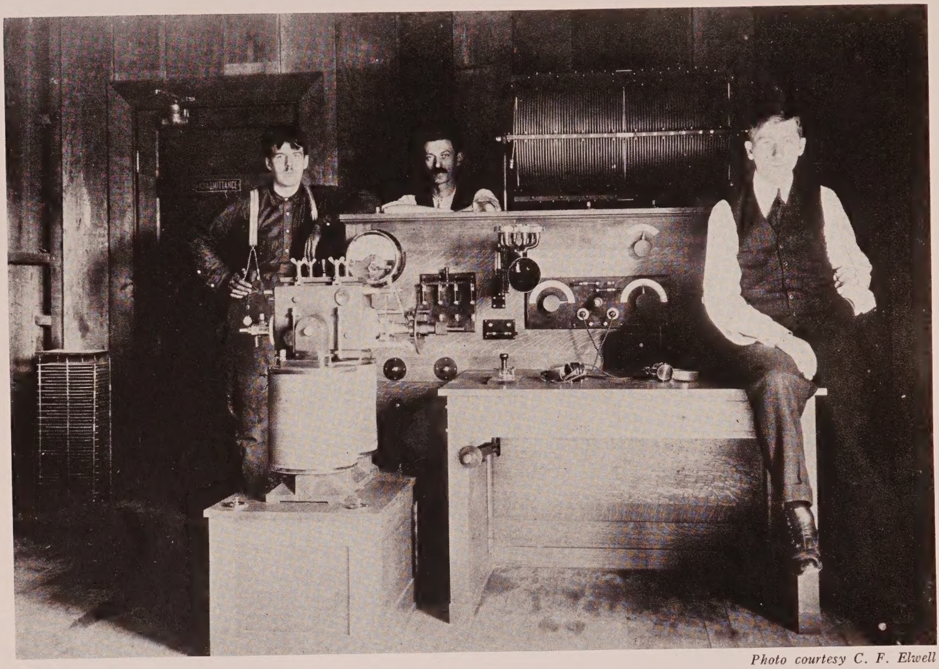


Photo courtesy C. F. Elwell

The first complete arc transmitter and receiver built by Federal. The microphone for radiotelephony with its many "buttons" to permit handling heavy currents may be seen in the center of the rear panel. From left to right, are D. Perham, C. Albertus, and Peter V. Jensen. Jensen later became well known as a manufacturer of loudspeakers.

While the two new stations were being tested, the operators of a large spark station located in Sacramento attempted to break up the tests by creating interference. This practice was not unusual in the days before radio laws were enacted in the U. S. A. and was accomplished simply by turning the transmitter on and letting it run or by "putting a brick on the key." The broad-tuning spark transmitters blanketed all reception in the vicinity and many were the rival radio stations put out of business by this process. At least these operators recognized the arc stations as potential rivals, a rivalry that was to continue for years after between the arc and the spark. The interference from the spark station did break up telephone tests, because the conventional detector and receiver were used for radiotelephone reception, but it had little or no effect on arc telegraph signals since a new type of detector, called a ticker, was used for reception of the undamped waves from the arc.

Dr. David Starr Jordan Officiates at Opening

The two stations were opened officially on February 19, 1910, with the press and public invited to attend the ceremonies. Dr. David Starr Jordan, president of Stanford University, had consented to go to Sacramento to officiate at the opening ceremonies. The following excerpt from an article which appeared in the *Sacramento Sunday News* February 20, 1910, gives some idea of the reception the press gave this opening:

> David Starr Jordan, eminent scientist as he is, probably enjoyed an experience yesterday in this city that had never been his lot before.
>
> He went out to Elmhurst and conversed through the wireless telephone station there with a Stockton man fifty miles away. Dr. Jordan freely expressed his gratification over the experience.
>
> The instrument certainly ranks with the world's greatest achievements. The instrument somewhat resembles the ordinary telephone, but is much larger. A Sunday News representative was among the considerable number who went out yesterday to test the new discovery by the practical method of holding conversation through it. The response from Stockton where the receiving end is at present located was rendered in a far clearer voice than the wire telephone is capable of. The tone was wholly free of the

metallic sound common with the regulation instrument. The voice was startingly natural. The Stockton man rendered Yankee Doodle with such effect as might have been expected had he been present on the spot.

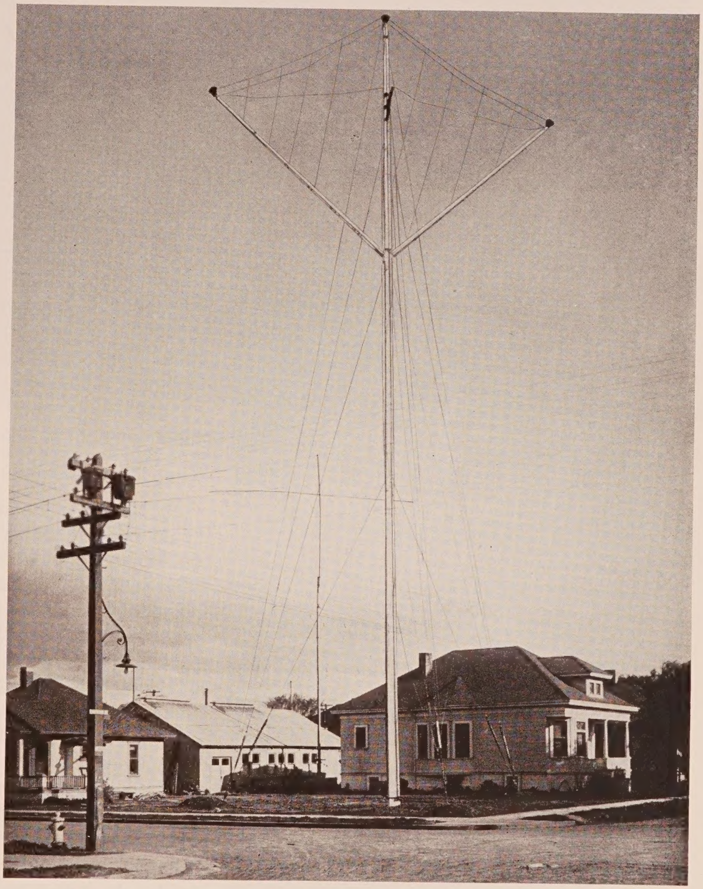
The capability of the invention is not limited to the mere transmission of sound. By a simple contrivance the telephone will carry three hundred words or more per minute and reproduce them faithfully at the receiving end.

The wireless telephone is free of the impediments that have been observed in the wireless telegraph that has been exhibited here. It is noiseless, to begin with, and there is no delay in reaching the receiver of any message desired to be sent. In point of speed and dependability the regulation telephone has no advantage whatever over the wireless, while the latter excels the old immeasurably in numerous respects.

Although sales of shares in the company, following the opening of the two stations, were good, the amount of each sale was small. It was soon realized that more substantial capital would have to be attracted to provide the funds for the patents and to permit the company to expand. However, there were still doubts expressed about the new system. One of the chief difficulties experienced with spark transmitters was their broad tuning and resultant interference when more than one station was operating at the same time. With only two arc stations, the sharp-tuning feature of undamped-wave transmission, a feature which finally delivered the death blow to damped-wave transmission, could not be demonstrated. It was decided, therefore, to build a third station at San Francisco in order to show that a number of continuous-wave stations could operate simultaneously without interference.

A Laboratory in a Shed

Since a third arc transmitter would be needed for the new station, the company decided to begin manufacture of its own arcs. Property was purchased at the corner of Emerson Street and Channing Avenue in Palo Alto. The new property had a small house near the corner and a shed in back of it fifteen feet wide by thirty feet long. This shed was doubled in size and became the first factory for the construction of Poulsen arcs in the U. S. A. This same shed later took on further historical significance since it was there



Federal's first laboratory and factory located on Emerson Street and Channing Avenue, Palo Alto, California. The first Federal arcs were manufactured in the corrugated iron shed behind the cottage, which was used as a laboratory. The shed became a place of historical significance when, in 1912, Dr. Lee de Forest and his assistants developed the audion amplifier and vacuum tube oscillator there.

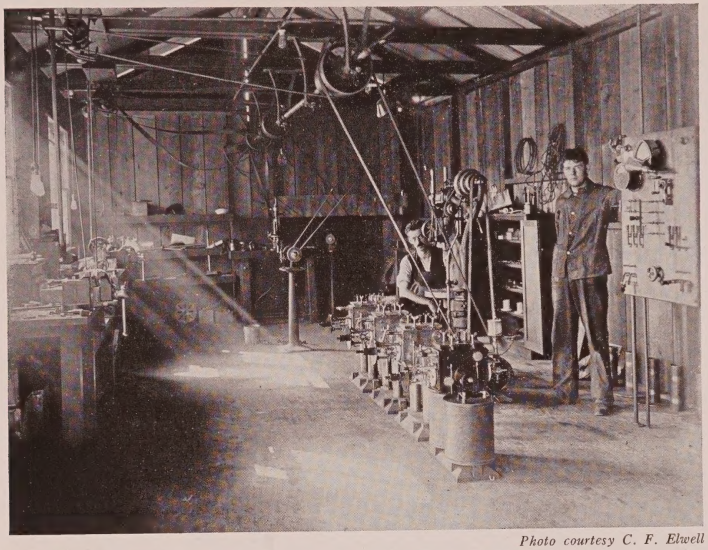
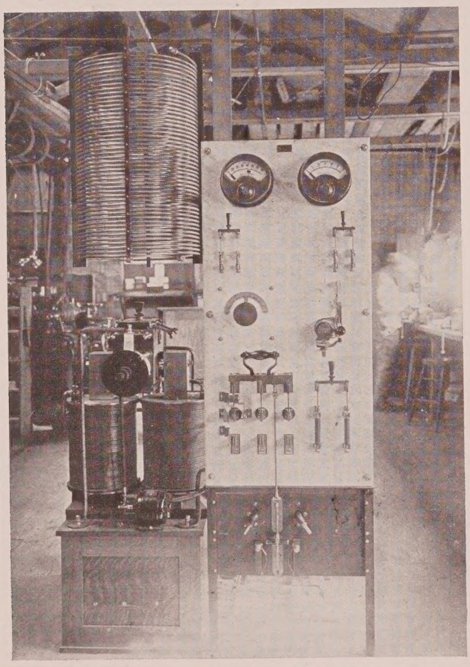


Photo courtesy C. F. Elwell

The first 5-kilowatt Poulsen arcs built in the U.S.A. With new design features incorporated by Federal, these four arc converters were the initial products of the Federal factory, which then occupied a space 15 by 30 feet. Kneeling behind the arcs is C. Albertus, Danish engineer, and to the right is D. Perham, a Federal employee.

that Dr. Lee de Forest and his assistants discovered the audion amplifier and the vacuum tube oscillator.

Four 5-kilowatt arcs were designed and built at the new plant. These arcs were an improvement over the Danish arcs since they were designed for continuous operation. The Danish arcs would heat up after a short period of operation, making it necessary to turn them off for a cooling period. But the arcs built at Palo Alto were equipped with water-cooling and other devices to make them more rugged, foolproof, and efficient. While the arcs were being built, auxiliary apparatus for the transmitters as well as receivers had to be designed and constructed so that the new factory became a complete plant capable of turning out fully equipped radio stations.



Complete 12-kilowatt Federal arc transmitter set up for test at Federal's first factory in Palo Alto, California.

The "Beach Station" at San Francisco

By this time, it was evident that the stations being built could be used for the commercial transmission of messages as well as for demonstrations to prospective investors. With Los Angeles, 340 air miles from San Francisco, already chosen as the site for the fourth station, it was decided to build a more powerful transmitter in San Francisco. A lease was taken on a block of land near Ocean Beach, San Francisco, and on this sand dune two wooden lattice masts, each 300 feet high, were erected. In order to save money, old lengths of cable which had served to propel cable cars along Geary Street in San Francisco were purchased and used to guy the 300-foot masts and support them against the Pacific Ocean gales that often blow along the beach.

This familiar "Beach Station" has since bowed its proud wooden latticework to the progress of

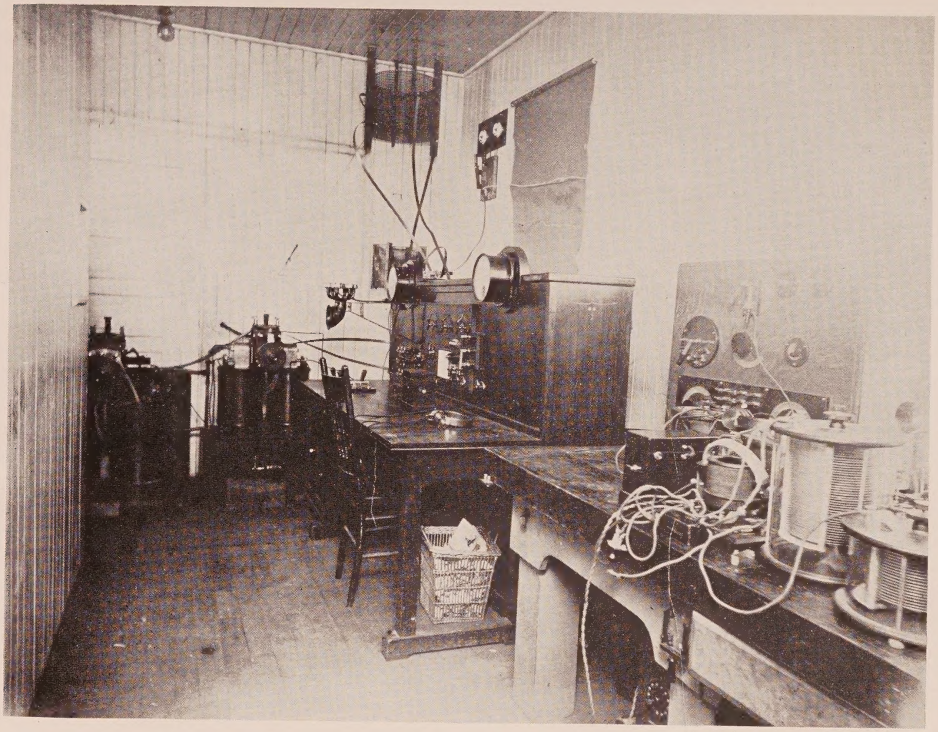


Photo courtesy C. F. Elwell

Interior view of the San Francisco "Beach Station" as it looked in 1909. This photograph shows a side view of the large and rather complex special carbon microphone used to modulate the arc transmitter for radiotelephony.

science. In its day it was a welcoming sight to incoming ships, whose masters used the towers to obtain a "fix" of their position. It also was the first commercial station to upset the notion that radio signals could only be efficiently received and transmitted high above sea level.

In order to cover the distance to Los Angeles, it was also decided to use 12- instead of 5-kilowatt arcs. So the four 5-kilowatt arcs built at the Palo Alto plant were installed in the Stockton and Sacramento stations, in duplicate, and 12-kilowatt arcs were built for San Francisco. It became the practice from then on always to provide a station with duplicate arcs so that one could be cleaned and adjusted while the other was in use.

With the third station in operation, it was possible to prove to skeptics that three stations could transmit simultaneously without interfering

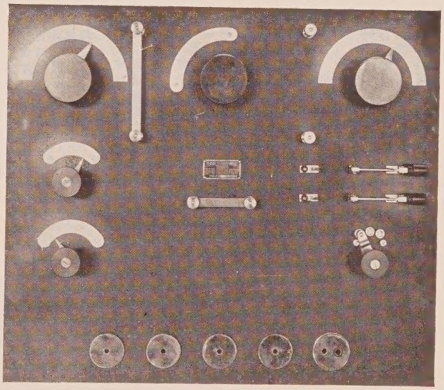


Photo courtesy C. F. Elwell

One of the earliest receiving cabinets built by the Federal Telegraph Company in 1911. The small disks at the bottom were jacks to permit telephones, detector, and ticker to be plugged in.

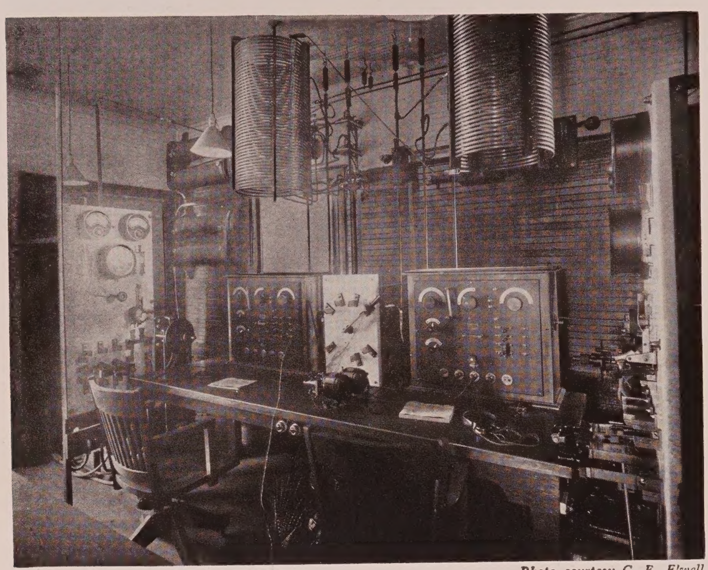


Photo courtesy C. F. Elwell

The Federal Telegraph Company South San Francisco station in 1913. This station was equipped for duplex operation so that two operators could handle messages at the same time. The transmitters were 30-kilowatt arcs.

with one another. It was easy to show that by simply turning a knob a very small amount, Stockton or Sacramento could be heard at will.

The San Francisco station was placed in operation in July, 1910. The demonstrations brought Beach Thompson, another Stanford graduate, to the station. Thompson had recently made a reputation for himself on the West Coast by organizing and financing the establishment of a new electric power line into San Francisco. He became interested in the system and proposed the formation of a new company with a recapitalization in order to provide working capital for further development.

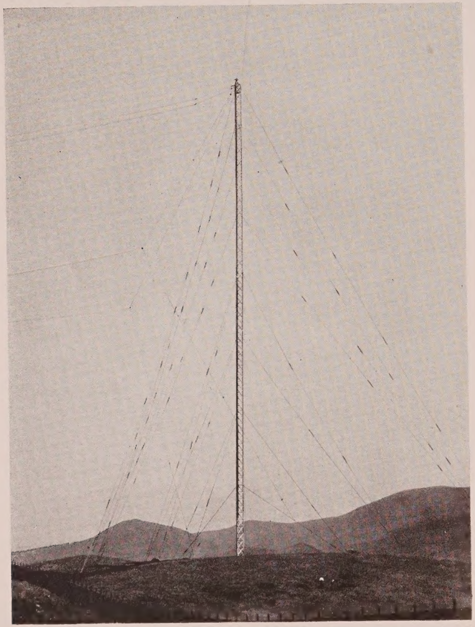


Photo courtesy C. F. Elwell

One of two 606-foot wooden lattice masts erected at South San Francisco in 1913 to permit reliable communication with the Federal station in Honolulu, where two similar masts were in service. These were the tallest wooden lattice masts ever built.

Poulsen Wireless Corporation

In January, 1911, therefore, the Poulsen Wireless Telephone and Telegraph Company was reorganized and the name was shortened to the Poulsen Wireless Corporation with Beach Thompson as president. Among the principal investors were E. W. Hopkins, George A. Pope, and J. Henry Meyer, all of San Francisco. Poulsen was paid $500,000 in cash and bonds in the new company for exclusive rights in the United States, including Alaska, Porto Rico, Hawaii, the Philippines, and Cuba. The original board of directors of the new enterprise included the names of Charles D. Marx, S. E. Slade, and Charles R. Bishop.

Almost from the inception of the old Poulsen Wireless Telephone and Telegraph Company, there had been a company set up as a wholly-owned stock company with a capitalization of only $100,000. Known first as the Wireless Development Company and later as the Federal Telegraph Company, this company had been organized for the purpose of avoiding high taxation in case business was done with states that set

Photo courtesy C. F. Elwell

Interior of the rapid telegraph room at the South San Francisco station. Traffic was handled here at speeds up to 180 words a minute, almost an unheard of speed at that time and a rate not often exceeded today.

their tax rate on the capitalization of the company. Since little or no business had been transacted outside the state of California, the Federal name had not been heard of until the reorganization. With the Poulsen Wireless Company as the holding organization and Federal Telegraph Company as the operating company, all business was transacted through Federal and the Federal name soon became known throughout the world. Poulsen arcs were shortly being called Federal-Poulsen arcs and a few years later Federal arcs.

The first statement issued by the board of directors of the new company was thoroughly conservative and hardly forecast a great future:

> The system has been demonstrated by us to be reliable for distances up to 750 miles at any time of day and night in any kind of weather and over land or water. Devices demonstrated by us in the laboratory, but not yet in the field, give promise of far greater distances. Dr. Poulsen, according to our latest advices, is now sending as high as 245 words a minute over land for distances over 900 miles, at any time of day. We have sent 300 words a minute from San Francisco to Stockton, but have not tried the rapid devices for greater distances. We are now talking by wireless telephone between San Francisco, Stockton, and Sacramento, but have not yet installed telephones at greater distances. Dr. Poulsen states, however, that he is able to hear the human voice clearly and distinctly for a distance of 420 miles.

Beach Thompson continued as president of the company until his death some four years later, and in 1915 George A. Pope became the second president.

More New Stations

The Los Angeles station was built in 1911 and Federal opened offices downtown, both there and in San Francisco, to take public messages between the two cities. The first day's income was a little over five dollars, but it increased steadily daily thereafter. Since this service was in direct competition with the two wire telegraph companies then operating on the west coast, a special rate was established to attract the public. This rate was set as the same for fifteen words and address as the wire companies charged for ten words and address. Also, 50 words and address were accepted for one and a half times the regular rate for ten words for transmission any time during the day at the company's convenience. This was the first day letter rate. Night letters carried 75 words and address for the same price as ten-word telegrams.

Because of the success of this communications business, it was decided to push service to the north and to the east. Stations were built in rapid succession at Portland, Seattle, El Paso, Fort Worth, Kansas City, and Chicago. In Chicago, the 26-story Transportation Building was used to support the station's 80-foot steel masts and the transmitter was installed on the twenty-second floor.

Atmospherics caused a great deal of trouble, especially in the daytime. To improve the speed and reliability of daytime service, two relay stations were built, one at Medford, Oregon, and one at Phoenix, Arizona. These stations also provided additional centers for business. At night, it was usually possible to skip the relay stations and work longer distances directly, such as between San Francisco and Kansas City. By 1912, 14 of these stations in the U. S. A. were competing with the wire telegraph companies. They communicated regularly over distances of 500 miles in the daytime and 1000 miles at night. Those were great distances then and demonstrated decided progress.

The Honolulu Station

At this time, the cable rate from San Francisco to Honolulu and Manila was 35 cents and $1.08 a word, respectively. It was determined that

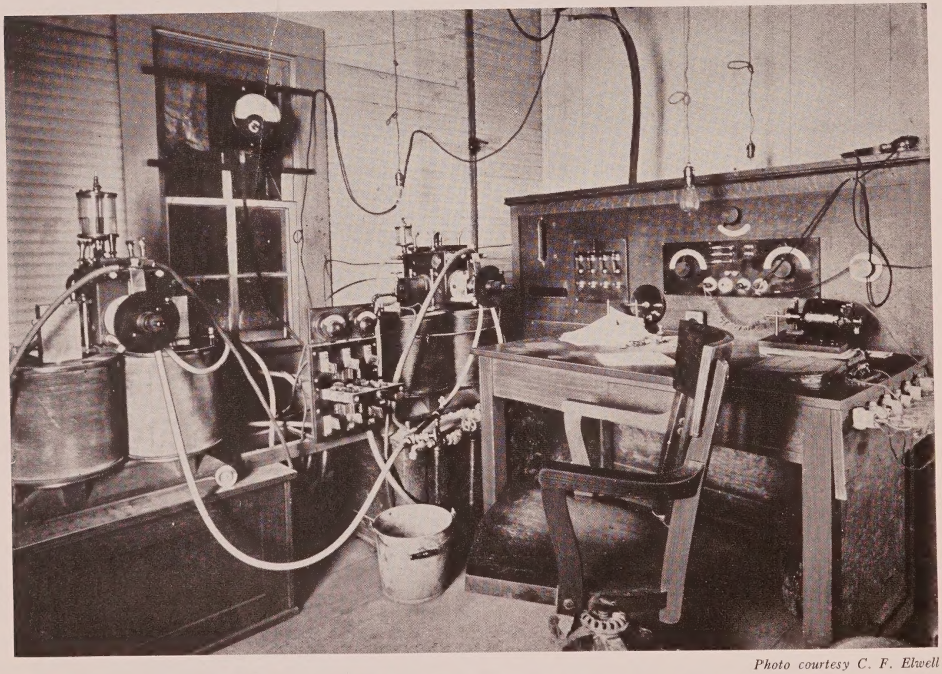


Photo courtesy C. F. Elwell

Operating room of the Federal 12-kilowatt station at Fort Worth, Texas, as it looked in 1912. Although the operating table closely resembles the first one built by Federal, no large-sized microphone was included as the original application of the arc to telephony had been superseded by the more pressing commercial demands for telegraphic communication.

actual cost of transmission between Honolulu and San Francisco would not be much greater than, say, between Chicago and San Francisco, for which an average of three cents a word was received. The board of directors of Federal, presented with these facts, provided the funds to build a Honolulu station and also to construct a larger station at San Francisco. Work on this project was expedited when it was discovered that the rival radio company, still using spark transmitters, was planning to build a big station in Honolulu.

As soon as this last fact was known, Federal lost no time in getting its chief engineer to Honolulu to erect the new station there. In fact, since no cabin was available on the next boat, he was accomodated in the officers' quarters. This great haste was reasonable enough, because in those days, once a high-power radio station was built, no one else would be foolish enough to build another station within many miles of it.

Federal had not previously erected any masts higher than 300 feet nor employed power greater than 12 kilowatts. At that time it was understood that the radiated power increased with the square of the height of the antenna and also with the square of the number of antenna amperes. So it was decided to quadruple the radiated energy by raising the mast height to 440 feet and by increasing power to 30 kilowatts.

A 2400-Mile Station in 47 Days

With the aid of two riggers from San Francisco and 25 Chinese and 25 native laborers, the Honolulu station was completed in May, 1912, 47 days after work was started. The new, more-powerful San Francisco station, located at South San Francisco in order not to be too close to the old station at Ocean Beach, had been placed in operation before the Honolulu station was started. Night communication between the two stations, a distance of approximately 2400 miles, was good, but daylight signals were not always reliable because of the prevalence of atmospherics.

An office was opened in Honolulu in competition with the cable company and the 35 cent-per-word cable rate was reduced to 25 cents. The press rate of 16 cents a word was reduced to two cents a word and a contract entered into with the local Honolulu newspaper for a minimum of 1500 words a day. Previously, 120 words daily was the maximum of outside world's news received by the islanders.

Photo courtesy C. F. Elwell

The Chicago station of the Federal Telegraph Company was located on the 26th floor of the Transportation Building. Two steel masts, illustrated in the above photograph taken in 1913, were built atop the building to support the antenna system.

To increase the strength of the daylight signals and improve reliability, 606-foot masts were later erected at both the Honolulu and San Francisco stations, and power was raised to 60 kilowatts.

Dr. Lee de Forest Heads Research

While these stations were being set up and more and larger arcs were being designed and installed, other developments took place in the tiny laboratory in Palo Alto that were destined to have a sweeping effect on wire telephony as well as radio communications. In July, 1911, Lee de Forest, inventor of the three-element vacuum tube he called the audion, became head of the research laboratory. A radiotelephone company de Forest had organized in San Francisco had failed and Federal bought up some of de Forest's equipment and made him a member of the laboratory staff.

Dr. de Forest had been employed specifically to attempt to develop a satisfactory vacuum tube detector for arc reception. The old-fashioned ticker was still in use then and it was never a too-efficient instrument. The first audions were purchased from MacCandless of New York, at the time the only manufacturer of three-element vacuum tubes. Later, a firm in Oakland began their manufacture to de Forest's specifications.

C. V. Logwood—the Practical Experimenter

When de Forest was made head of research at the Federal Laboratory, he was assigned a young man by the name of C. V. Logwood as his assistant. Logwood had been an amateur radio experimenter. He had not had much technical education, but he was a born experimenter. His first achievement at the laboratory was to

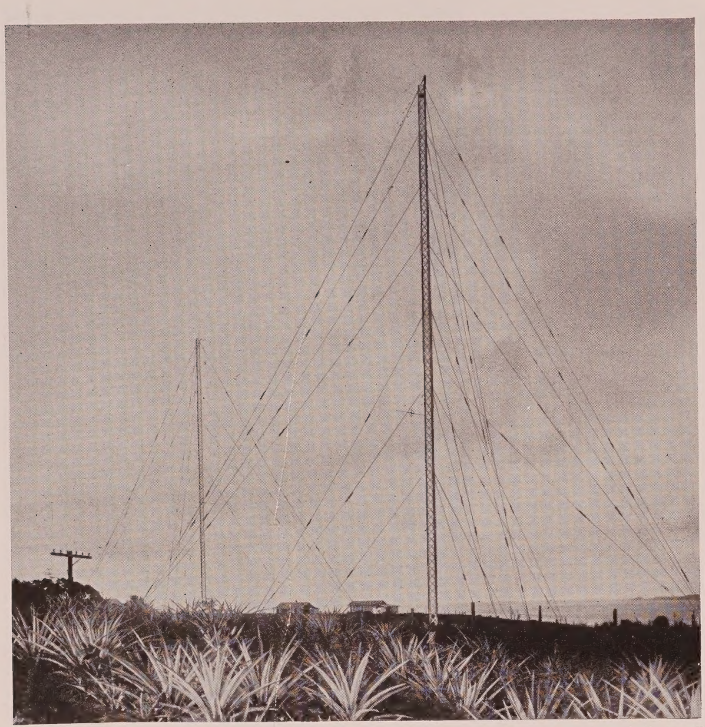


Photo courtesy C. F. Elwell

Federal Honolulu station showing buildings and the two 606-foot wooden lattice masts as they appeared in August, 1912.

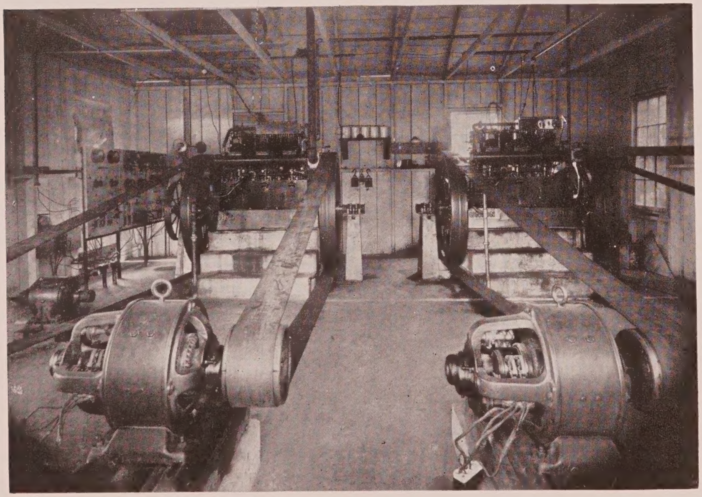


Photo courtesy C. F. Elwell

Power plant for the Honolulu station showing the relatively huge gasoline engines used to drive the direct-current generators for the arc converters.

design a rotary ticker-detector the cost of which was a fraction of that of the elaborate, gold-wire, vibrating type brought over from Denmark. Logwood simply roughened the pulley of a $1.25 toy electric motor and arranged a small piece of steel wire so that it bore on the roughened groove. The result was a ticker which gave stronger and better sounding signals than the much more expensive type then in use.

At first the differences between the two men—de Forest a doctor of philosophy from Yale, and Logwood a practical experimenter of the type so common in America, but with little formal education—caused some friction. These differences soon ironed themselves out, however, and together, in the Federal laboratory, they made two of the most important discoveries of their time and, even after they left Federal, they worked together and filed numerous joint patents.

Invention of the Audion Amplifier

In August, 1912, Logwood, while experimenting with the de Forest audion, hooked three stock Federal PN audion detectors in cascade, each connected to separate filament and plate batteries. With a microphone in the input of the first unit and a pair of headphones hooked up to the last unit, the ticks of a watch placed in front of the microphone were amplified so that the headphones reproduced them loud enough to be heard all over the room. It was even possible to hear a fly walking on a piece of drawing paper placed before the microphone. This was the first application of the audion, or three-element vacuum tube, as an amplifier.

In September, 1912, Federal sent the first self-contained three-stage amplifier to Washington, D. C., to demonstrate it to the U. S. Navy. This amplifier incorporated the three audions in a single box with suitable switches, resistors, transformers, etc., with the plate battery inside and with terminals for a single external battery to light the filaments. Dr. Louis Austin, then head of the U. S. Naval Radiotelegraphic Laboratory, took the amplifier to the Bureau of Standards and measured its amplification. With the crude methods available at the time, amplification was calculated to be 120 times.

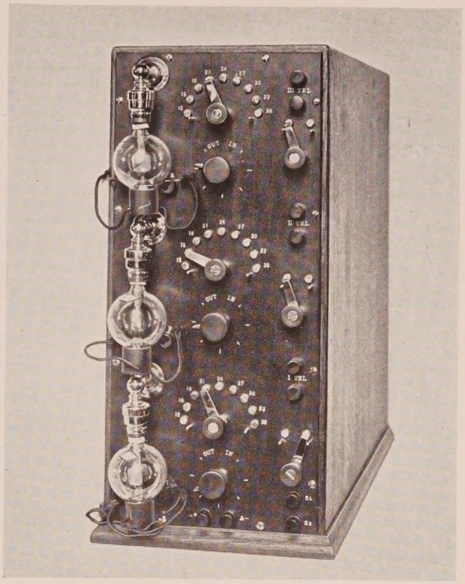


Photo courtesy C. F. Elwell

Three-stage de Forest audion amplifier of the type first built by Federal in 1912. This is the earliest known commercial cascade audio-frequency amplifier and is the type demonstrated to the U.S. Navy in September, 1912. At that time, tests made by Dr. Louis Austin, then head of the U.S. Naval Radiotelegraphic Laboratory, indicated the amplifier increased the signal intensity by 120 times.

Discovery of the Audion Oscillator

While experimentation was progressing on the amplifier, another engineer by the name of van Etten, who was working with de Forest and Logwood, noticed that the audions had an annoying tendency to produce unwanted squeals in the headphones. Further investigation of the phenomenon by de Forest revealed that the audions were oscillating and producing not only audible squeals but inaudible radio frequencies as well. In fact, when an audion was hooked up in the laboratory with capacitance and inductance like a Poulsen arc, it generated undamped oscillations identical to the oscillations from the arc. The audion oscillator, moreover, had two important advantages over the arc. It could be made to operate on power down to a fraction of a watt while the smallest Poulsen arc capable of satisfactory operation consumed at least 100 watts. Also, the audion could be operated at much higher frequencies than the arc and with greater stability.

In a sense, then, it may be said that the arc actually spawned its own successor. For, while arc transmission enjoyed a position of preeminence in radio for over twenty years, it has now been replaced by the vacuum tube transmitter. Actually, this accidental discovery of the vacuum tube as an oscillator and the invention of the vacuum tube amplifier may be said to have started a whole new division of science now known as electronics, since most vacuum tubes in use today are employed either as oscillators or amplifiers. Fortunately, van Etten kept notes on the experiments; for, when the whole question of patent rights to these inventions came up later in the courts, it was his notes that established the fact that the original work was done in the Federal Telegraph Company's Palo Alto laboratories in August, 1912.

Audion Amplifier Aids Telephony

In October, 1912, de Forest took the audion amplifier to New York and demonstrated it to officials of the American Telephone and Telegraph Company. Telephone engineers had been seeking such a device for years to extend the distance a telephone conversation could travel over open-wire lines. The limit then was about 900 miles. Early in 1913 the Telephone Company bought the wire telephone rights to the amplifier from de Forest and later it purchased the telegraph rights. In subsequent litigation, the courts decided that Federal was to retain the shop rights in this invention because of the development work done in its laboratory.

The fact that Federal kept the shop rights in these basic de Forest patents was of great importance to the company after World War I when the vacuum tube became so important in radio broadcasting, high-power transmission on short wavelengths, and the construction of amplifiers for loudspeaker operation. When the Radio Corporation of America was formed shortly after the war for the purpose of pooling the patents of a number of communications companies, the American Telephone and Telegraph Company turned the de Forest patents into the Radio Corporation pool. The Federal Telegraph Company was, therefore, for a number of years the only company in the United States outside of the Radio Corporation group in a position to supply equipment in all fields of radio and other phases of electrical communication based on these fundamental and controlling de Forest patents.

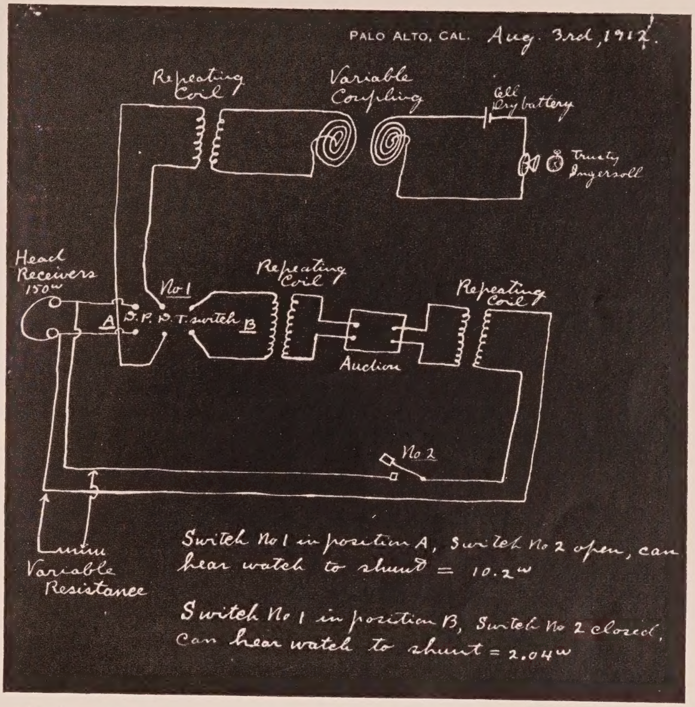
U.S. Navy Adopts Federal Arcs

The year 1912 was an eventful one in Federal's history. That was the year when the company had established itself in the communications business with radio stations as far east as Chicago and with a station in Honolulu. It was the year when the laboratory in Palo Alto was the scene of two epoch-making developments in radio: the audion oscillator and the audion amplifier. And 1912 was also the year when the United States Navy officially adopted the Federal arc.

When Federal's chief engineer took the audion amplifier to Washington in September, 1912, he saw Admiral Hutch I. Cone and requested permission to demonstrate a Federal arc to the Navy Department. The Navy had then just completed its new 100-kilowatt spark station at Arlington, Virginia. This station had taken three years to build and was considered the last word in radio equipment with its one 600-foot and two 450-foot self-supporting steel antenna towers. The navy had tested all the various types of spark systems then available and had finally decided on the 100-kilowatt Fessenden transmitter.

Had Federal not successfully established its 2400-mile circuit between Honolulu and San Francisco, a prestige-making feat in days when a jump of 1000 miles was considered unusual, the Navy might not have considered the arc. After all, the Navy officials were naturally very pleased with their brand-new 100-kilowatt Arlington spark station which they considered the most up-to-date high-powered radio equipment available. As it was, Admiral Cone reluctantly granted permission to a test of the arc at Arlington, but he also issued a proviso "not to put any nails or screws in the floors, walls, or ceilings of the station." Dr. Austin was openly skeptical of the performance of the arc when he heard that the 30-kilowatt transmitter planned for the demonstration would put only about fifty amperes into the antenna. "What is the use of fifty," asked Dr. Austin, "We have over a hundred." When it was pointed out that Federal considered an ampere of continuous waves worth two amperes of damped waves, Austin replied, "Rubbish, an ampere is an ampere the world over."

To comply with Admiral Cone's injunction not to mar the new Arlington station in any way, a wooden frame eight feet long, eight feet high, and four feet deep was built. In this frame was suspended all the apparatus necessary for the 30-kilowatt arc. Next to the huge Fessenden spark transmitter, the arc transmitter looked small and ridiculously uncomplicated.



PALO ALTO, CAL. Aug. 3rd, 1912.

Switch No 1 in position A, Switch No 2 open, can hear watch to shunt = 10.2w

Switch No 1 in position B, Switch No 2 closed, Can hear watch to shunt = 2.04w

Circuit of the first experimental audion amplifier as developed under the direction of Dr. Lee de Forest in Federal's Palo Alto laboratory in 1912. This diagram served as an important document when the question of patent rights to the invention of the amplifier came up later in the courts. It helped establish the fact that the prior work was done in Federal's laboratory which subsequently won for Federal the shop rights to the invention.

Arlington to Honolulu in One Hop

The day the arc was installed (December 8, 1912), communication was established between Arlington and Federal's South San Francisco station. This was considered phenomenal by officials of the Navy since, up to that time, the Navy had not succeeded in contacting its Mare Island station in California, almost the same distance. The following night, messages were exchanged between Arlington and Federal's Honolulu station, a distance of 4500 miles and a record for that time, especially since, during the period of transmission, Honolulu was still in daylight.

Arc Proved Superior to Spark

Dr. Austin insisted that more scientific tests be arranged to compare the relative performances of the 30-kilowatt arc and the 100-kilowatt spark, so the arc was rented by the Navy for further study. He first arranged audibility tests at Key West, Florida, where, in spite of the fact that the arc was radiating only one-quarter of the energy transmitted by the spark, signal strengths of both transmitters were equal. Later tests at Colon, Panama Canal Zone, yielded significant results inasmuch as the messages sent with the arc were received through the heavy tropical static while those from the spark were not. During this period, the cruiser *Salem* travelled as far as Gibraltar in order to compare the two systems thoroughly.

While tests on the arc were in progress, Dr. Austin also had an opportunity to make comparative tests of Federal's receiving equipment against the receiver then in use at Arlington. The official Navy receiver at that time was the IP76 and Naval radio experts considered it the last word in radio receivers. However, in audibility tests, Dr. Austin found that the Federal receiver provided from 2.8 to 4.6 times as much energy as the IP76. Thus, Federal not only had a superior transmitter in the arc, but it also proved to be ahead in the design of receiving sets.

U.S. Navy Installs High-Power Federal Arcs

Following these tests, the United States Navy ordered a 100-kilowatt arc for installation at Darien, Panama Canal Zone. Along with the arc transmitter, the Navy also purchased Federal ticker receivers, equipped with oscillating audions for the shorter-wave work. This equipment was installed in 1913, thus making the United States Navy the first in the world to adopt continuous-wave transmission. In quick succession, the Navy purchased 350-kilowatt arc transmitters for stations at Pearl Harbor, Hawaii, and Cavite, Philippine Islands. The Federal arc converters for both these stations were of the oil-immersed, water-cooled type capable of furnishing 350-ampere antenna current continuously. These arcs looked like small vertical power-station generators. They were nine feet two and a half inches tall, seven feet four inches wide, and twelve feet long. Each magnetic field weighed 65 tons. A smaller station of 200 kilowatts capacity was also installed at San Diego, California, to work with the two Pacific stations. Some idea of the confidence the Navy had in the ability of the arc to work long distances may be gathered from the mileages between the stations. The distance from Cavite to Pearl Harbor is 5300 miles, from Pearl Harbor to San Diego 2600 miles, and from Cavite to San Diego 7800 miles.

Federal Inaugurates Ship-to-Shore Service

In 1914, the Federal Telegraph Company entered the marine radio field by inaugurating a ship-to-shore service and manufacturing radio equipment for installation on ships. This service was introduced at San Francisco and, for the first time, ships plying between San Francisco and Honolulu were able to secure daylight communication practically anywhere between these points. In 1921 and 1922 the service was extended by the establishment of coastal stations at Los Angeles, California, and Portland, Oregon.

During the years that followed the adoption of the arc by the Navy until the United States entered World War I in 1917, the Federal Telegraph Company devoted most of its manufacturing efforts to providing radio equipment for the United States Navy. Since so many of the transmitters were 60 kilowatts or over with arc converters weighing as much as 85 tons, the plant was moved to a new site in Palo Alto close to a railroad siding. The small group of employees had, by this time, grown from about a dozen to over 200.

A 1000-Kilowatt Arc Station

It was during this period that the U. S. Navy Department set up a comprehensive radio system all over the United States and its possessions. In addition to the four Navy stations previously mentioned Federal arc stations were built and placed in operation at Sayville, Long Island; Tuckerton, New Jersey; Annapolis, Maryland; Guam; and Porto Rico. After the United States entry into the war, the U. S. Navy ordered two 1000-kilowatt arc converters for installation at Bordeaux, France, with a view to supplementing the existing means of communication between France and the U. S. A. This was the famous Lafayette Station, the most powerful radio station ever built. Installation by Federal engineers was begun in 1918, but the Armistice halted work for a while until negotiations were made for the French Post Office to operate the huge station. It was finally completed and opened in December, 1920, and arcs were used until 1936 when the station was converted to vacuum tube operation.

Operating Innovations

By 1917 the Federal Telegraph Company's chain of radio stations along the Pacific Coast was doing such good business that duplex circuits were set up between Los Angeles and San Francisco. These circuits permitted transmission of two messages between two points simultaneously. Also, one-way "break" systems were set up between Los Angeles and San Diego, Los Angeles and Phoenix, and San Francisco and Honolulu. The break system permitted the receiving operator to break in on the transmitting operator whenever an error occurred or, for some other reason, continuity of reception was interrupted.

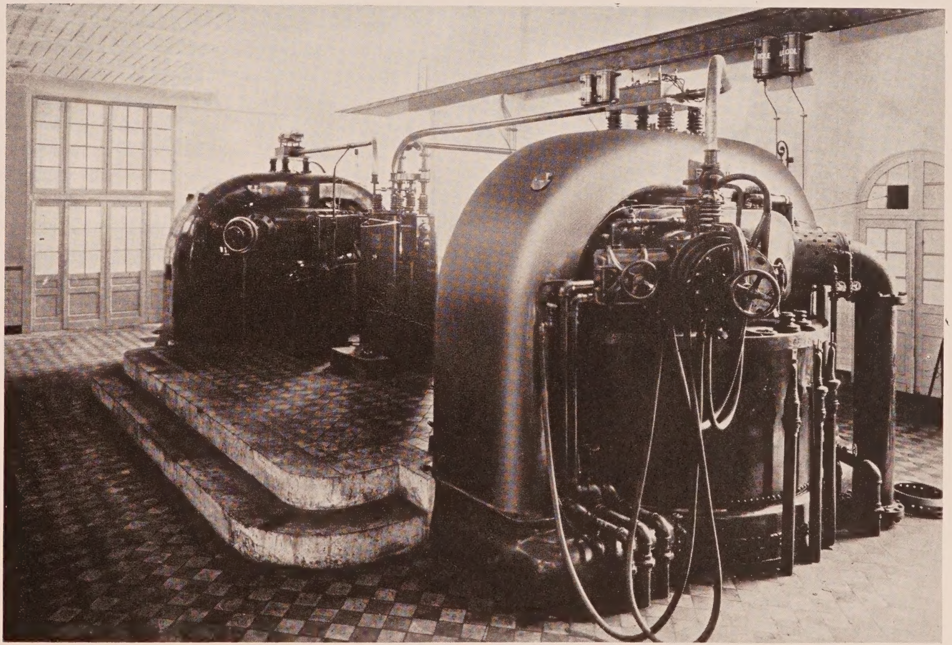
At most places, the Federal Telegraph Company receiving station was located at a central point in the city being served and the transmitter, usually on the outskirts of the city, was operated by remote control from the receiving office over leased wires.

Operating innovations such as duplexing and break systems were only a few of the many ingenious methods employed by Federal to simplify and speed the handling of messages. As early as 1913, Federal installed Wheatstone, tape-operated, Brown relay keying between San Francisco and Los Angeles. Messages could be sent at 80 words per minute and were received on the Telegraphone, another Poulsen invention, which permitted recording of sound on a rapidly moving steel wire. After the signals were recorded, the wire was run at a slower speed so that the operator could transcribe the message directly on his typewriter. The Federal audion amplifier made this development possible.

U.S. Navy Purchases Federal Radio Stations

When the United States entered World War I, the Navy for security purposes took over all of Federal's stations. Much of the equipment was subsequently transferred to military bases throughout the world and Federal engineers were drawn upon by the Navy to supervise the installation and operation of these stations.

The Government paid $1,600,000 in Liberty Bonds for the radio stations and virtually all of the company's domestic patents, the company retaining, however, the Palo Alto manufacturing and engineering unit and its foreign rights. With the money, a long-standing indebtedness amounting to about $600,000 was cleared up. At the time, Washington Dodge was president of the company, but shortly after he was succeeded by R. P. Schwerin. The new capital permitted the company to expand its manufacturing facilities, and, throughout the war, Federal was one of the



Historic Lafayette radio transmitter station built by the Federal Telegraph Company for the United States Navy near Bordeaux, France, during World War I—the largest transmitter ever built. The illustration shows the two 1000-kilowatt Poulsen arc generators which permitted efficient and reliable communication across the Atlantic Ocean.

principal suppliers of radio equipment for the Navy and Maritime services. With increased plant and personnel, Federal responded to the call by the mass production of radio equipment serving the Emergency Fleet—the equivalent of today's Victory Fleet. Thus it may be said that Federal's products contributed substantially to the supremacy of the Allied Nations in World War I.

After hostilities were ended, the Government restored to Federal all of its patent rights. A new system of radio stations was built under the direction of Haraden Pratt who was then acting chief engineer. Since the company also had managed to continue its commercial message service through its wire-leasing arrangement with the Pacific Telephone and Telegraph Company during the time that the Government operated Federal's radio stations, its position was better than at any previous time.

Frederick A. Kolster Heads Research

In 1921, Frederick A. Kolster joined Federal as chief research engineer. From 1912 to 1921 he had served as radio specialist at the United States Bureau of Standards and was chief of the radio section which he organized. While at the bureau, he attained recognition for his invention of the Kolster Decremeter, widely used by the Radio Inspection Service and others for determining the wavelength and logarithmic decrement of radio transmission, primarily from ships, as required by government regulations established for the purpose of promoting safety at sea.

Kolster also made another important contribution to safety at sea when, in 1913, he prevailed upon the Bureau of Lighthouses to establish radio beacons at all important lighthouses and lightships. As a result, the first experimental radio beacons were installed by Kolster on the Ambrose and Fire Island lightships and at the Sea Girt lighthouse on the New Jersey coast.

Kolster had been employed specifically for the commercial development of his radio compass or direction finder which had been used by the United States Navy during the first world war as one of that conflict's "secret weapons." Kolster's direction finder remedied the faults of older systems. The Kolster system for pick-up purposes utilized a multiple-turn loop or coil antenna which could be rotated about a vertical axis. In this way, by revolving the loop for maximum signal strength, bidirectional results could be obtained. Kolster also used a loop in conjunction with an antenna to obtain a unidirectional bearing. The latter was based on the maximum signal strength and was rather broad. The bidirectional signal gave the accuracy, and the unidirectional signal the information as to whether transmission was coming from front or rear. The unidirectional feature was an improvement over previous systems.

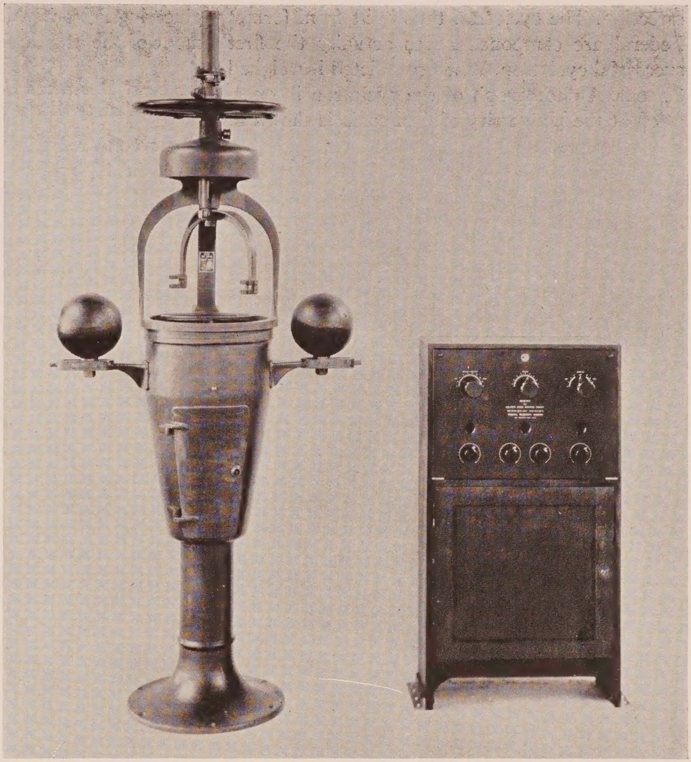
This new-type loop had greater pickup than previous systems and, when used with a vacuum tube receiver, its directional sensitivity was greatly improved. Installed on shipboard, the radio direction finder provided a means of determining the bearing of any radio beacon or of a radio-equipped ship lost in fog or in distress.

Federal Builds First Commercial Radio Direction Finder

Through Kolster's efforts, a commercial radio direction finder was developed and manufactured. This was the first radio direction finder to be installed on commercial ships. Federal thus introduced to the maritime world a simple and effective method of utilizing radio as a navigational aid. The radio compass or direction finder devised was, when it was introduced, described by the United States Commissioner of Lighthouses as the most important nautical invention since the magnetic compass. Today, a radio direction finder is an important piece of navigational equipment on almost every ocean-going vessel, and special radio compasses and direction finders developed and manufactured by Federal for the United States Navy were strategic "secret weapons" of World War II, just as the Kolster direction finders were in the previous war.

Contract with China

In January, 1921, the Federal Telegraph Company entered into a contract with the Republic of China for the construction of high-power radio stations at Shanghai, Harbin, Canton, and Peking for a contract price of $13,000,000. The original contracts provided for 1000-kilowatt and 60-kilowatt arc stations at Shanghai, 200 kilowatts at Harbin, and 60 kilowatts at Peking



Kolster radio position finder; the first practical radio direction finder for shipboard operation, developed by Frederick A. Kolster at the Federal Laboratories in 1921 and manufactured by Federal. At the left is shown the loop compensator mechanism as attached to the standard ship's binnacle. The rod at the top supported a square rotatable loop antenna located above deck. The vacuum tube receiver employed with the direction finder is shown at the right.

and Canton, but a supplementary contract, drawn up September 15, 1922, provided for secondary stations of 100 kilowatts capacity instead of the 60-kilowatt stations at Shanghai, Peking, and Canton, and a second 200-kilowatt station at Harbin. Japanese opposition, however, prevented completion of the contract.

The 1000-kilowatt arcs were to be constructed from castings made during World War I for a station to be built in North Carolina to communicate with the Lafayette Station, then under construction in France. The North Carolina project was cancelled with the signing of the Armistice on November 11, 1918, and the castings were stored at Federal until the China contract was negotiated, when actual assembly was begun on them. A photograph of the two castings nearing completion in the Federal factory at Palo Alto, California, is shown in the frontispiece. The photo was taken in 1921 before it became evident that the Chinese contract would not be consummated.

Federal Aids Cyclotron Development

These giant magnetic field pieces were stored at Palo Alto for a number of years while developments destined to have an important effect on World War II were taking place at the University of California in Berkeley. It was here that Dr. Ernest O. Lawrence was experimenting with early cyclotron models which later proved so important in the atom-smashing technique that led to the discovery of atomic energy secrets.

Dr. Lawrence's first models were tiny compared to the huge cyclotrons later successfully used to smash the atom. Since Federal was a pioneer in the electronic field, and the cyclotron was essentially an electronic device, Dr. Lawrence turned to Federal engineers in the mid-twenties for help in constructing the magnetic circuits for his preliminary experiments.

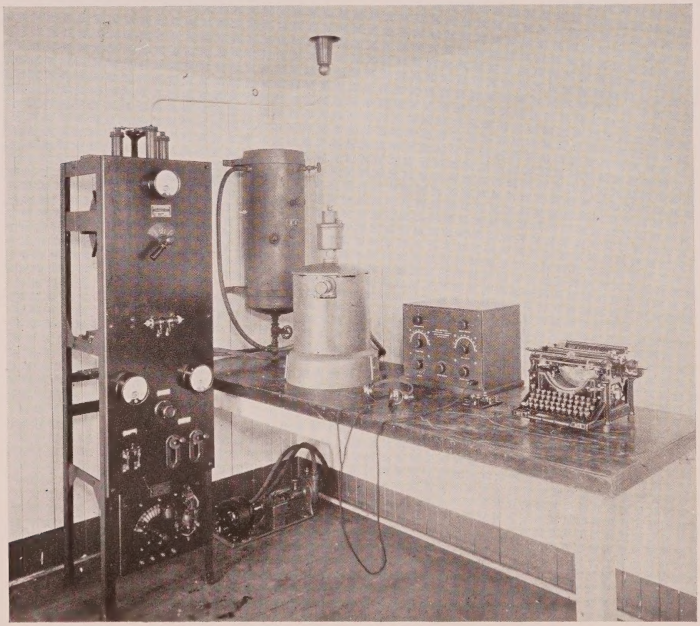
As a result of these first contacts, Federal was quick to appreciate Dr. Lawrence's problem in obtaining the massive magnets necessary for work on a larger scale and consequently donated the 65-ton castings intended for the China station to the University of California. Federal also wound the new field coils under Dr. Lawrence's direction. The cyclotron thus built from former Federal arc components was actually the first successful cyclotron. When completed it weighed 80 tons. A photograph of the cyclotron in operation at the University of California is shown in the frontispiece.

So perfect were the magnetic fields of Federal arcs for this purpose, that other old high-power arcs, when they were retired from use as radio transmitters by modern vacuum tube equipment, found their way to the physics laboratories of various universities where they made important contributions to the progress of science. Thus, the powerful arc transmitter built by Federal for the U. S. Naval Radio Station at Annapolis, Maryland, when it was dismantled after years of successful operation, was turned over to Dr. Joseph R. Dunning of Columbia University for the cyclotron constructed at the Michael Pupin Laboratory there. It was at this laboratory that the theories of Dr. Lise Meitner were first tested and verified by Professors Niels Bohr and Enrico Fermi. These experiments were the first step in the fabulous Manhattan Engineering Project which finally led to the development of the atomic fission bomb.

Federal Becomes Sole Operating and Controlling Company

A second reorganization of the company took place in 1920 when the stockholders ratified a plan of recapitalization to reduce a topheavy debt structure created in the original organization of the Poulsen Wireless Corporation in 1911. Under the presidency of R. P. Schwerin, the capital stock liability of the Poulsen Company was liquidated by exchanging Federal Company stock for Poulsen stock. The effect of this procedure was to produce a new capitalization of $2,500,000 and to eliminate completely the former Poulsen Wireless Corporation, the Federal Telegraph Company becoming the sole operating and controlling company.

In 1925, Federal entered the field of manufacturing and selling radio broadcast receiving sets which had been designed by its chief research engineer, Dr. Kolster. The new receivers were sold as Kolster Radios, and in a relatively short time they became the third largest factor in this branch of the radio manufacturing business in the United States.



Typical ship installation of a Federal Telegraph Company 2-kilowatt arc transmitter.

Federal Expands to East Coast

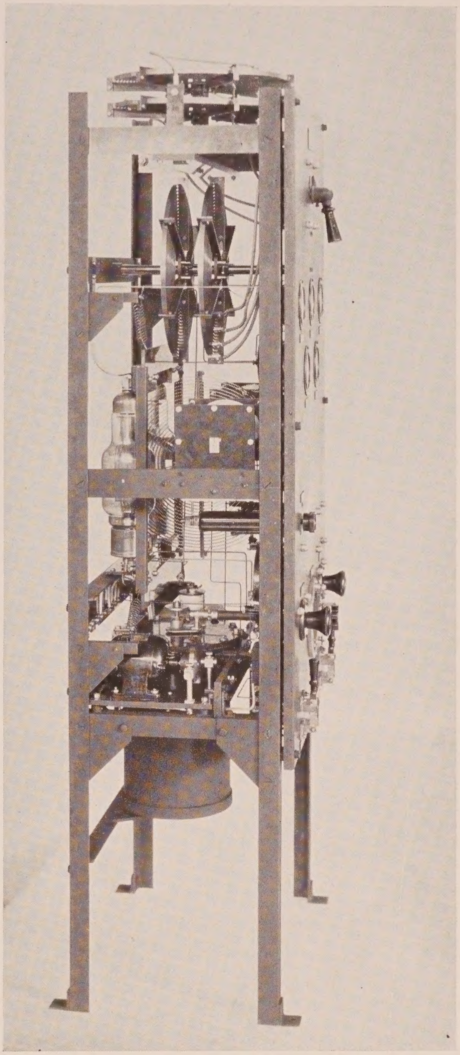
At the time of the introduction of the Kolster Radio receivers, the Federal plant in Palo Alto was in no way sufficiently equipped to meet the constantly developing demand. Federal, therefore, sought an eastern manufacturing organization capable of producing receivers in sufficient quantity with the quality called for by the high-class field to which the receivers catered.

This organization it found in the Brandes Companies of New Jersey, a group of manufacturing, engineering, and selling units which had been organized in 1908 and which enjoyed an enviable prestige in the radio trade. The Brandes business had been built up almost entirely around the radio headset which, with the development of the de Forest vacuum tube as an amplifier and the loudspeaker made possible by it, had rapidly declined in popularity, leaving the Brandes Companies in possession of large and fully developed manufacturing facilities and a vanishing volume of business.

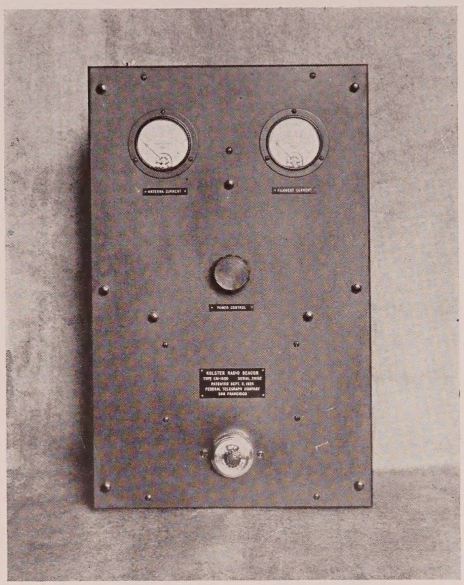
Manufacture of Kolster Radios was forthwith begun at the Brandes plant located in Newark, New Jersey, and a merger, based on an exchange of stock, was arranged. The new company formed, named Federal-Brandes, Inc., operated as a holding company for the Federal Telegraph Company and for the combined Brandes companies, thus effecting a union of two of the oldest radio companies in the country, having at its outset a fully developed business with manufacturing plants located at Newark, New Jersey; Palo Alto, California; Toronto, Canada; and Slough, England. In 1928, the name Federal-Brandes in the U.S.A. was changed to the Kolster Radio Corporation. By 1930 the status of radio set sales to the public had become so unstable, and so much of the plant capacity was being devoted to provide equipment for the expansion and modernization of the Mackay System, that it was decided to discontinue the manufacture of Kolster radios in the U.S.A. and to concentrate on the manufacture of vacuum tubes, radio transmitters, and navigational equipment.

Federal Joins Mackay System

In 1927, Federal took one of its most important steps forward when it completed a twenty-one year contract with the Postal Telegraph-Commercial Cables group, which comprised the Mackay System. By the terms of the new contract, the Federal Telegraph Company agreed to devote its patents, manufacturing, and engineering facilities to supplying the Mackay companies with whatever they needed to begin at once a land and ocean radio service on a parity with existing competition. As a first step toward aiding Mackay in the immediate inauguration of a radio service, Federal sold outright to that company its Pacific Coast point-to-point radio communication system, comprising the entire operating communication interests of the original Federal Telegraph Company. Under these arrangements the Federal Telegraph Company



Early 500-watt vacuum tube radiotelegraph transmitter manufactured by Federal in 1926.



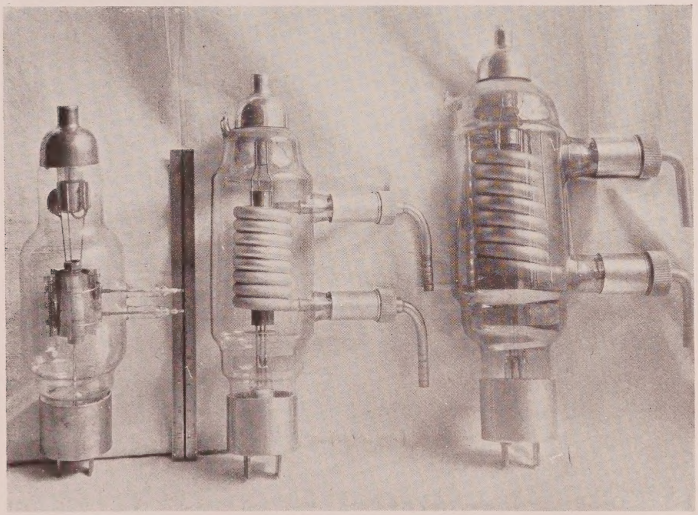
Radio Beacon Type CM-1530, developed by Frederick A. Kolster. It was manufactured by the Federal Telegraph Company in 1927.

became purely a manufacturing and engineering company engaged in the design and supply of electrical equipment to the Postal Telegraph–Commercial Cable Companies and their subsidiaries—now augmented by the Mackay Radio and Telegraph Company which was organized to take over the Pacific Coast Federal radio stations.

By this time the arc was fast being eclipsed by the vacuum tube as a generator of continuous waves, especially for the newer short-wave transmitters which could be built and operated so economically with beam-type antenna systems. In anticipation of this development, the Palo Alto plant was equipped in 1927 with its own vacuum tube manufacturing department. In the same year, the company also obtained the services and rights to patents of F. S. McCullough, one of the pioneer vacuum tube engineers in America for the development of this phase of the company's manufacturing business.

Kolster's Experiments with Directional Antennas

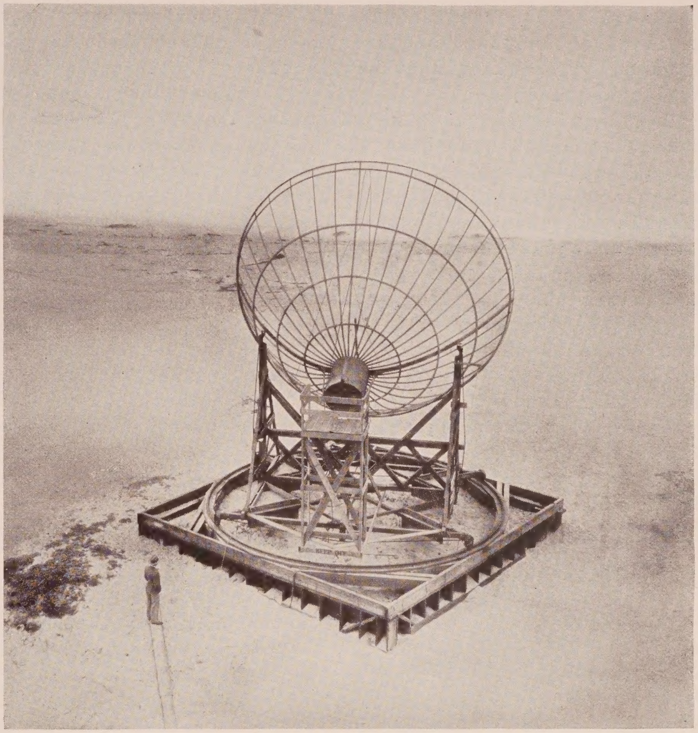
Further, early in 1928, Dr. Kolster began a series of experiments with directional antennas employing short waves. Investigation of the radiation patterns of many types of antenna systems was made to gather information regarding their directive characteristics. This work was carried on with equipment capable of operating on about three meters, considered at the time an extremely short wavelength for actual transmission, especially since transmission tests were made between Palo Alto and Seattle (710 miles) and Palo Alto and Honolulu (over 2000 miles). Kolster investigated three main types of antenna systems—parabolic



Three of the early type McCullough transmitting tubes manufactured by Federal in 1928. Left to right they are of 1/4-kilowatt, 5-kilowatt, and 10-kilowatt capacity. The two tubes on the right are equipped with water-cooled plates.

reflectors, loop antennas, and unidirectional antennas.

As a result of Kolster's experiments with directional antennas and the subsequent experience with high-frequency, high-power transmitting equipment, Federal was soon in a position to take the lead in creating for the Mackay System a point-to-point short-wave radio network utilizing vacuum tubes and transmitters of its own design and manufacture as well as directive antennas of superior efficiency.



One of the large parobolic reflectors used by Frederick A. Kolster in 1928 for a series of experiments with directional antenna systems employing short waves. These experiments led directly to later developments in microwave directional transmission.

Federal Becomes a Member of I.T.&T. System

The year 1928 saw another development of major significance in electrical communications in which Federal participated. On March 21, 1928, came the announcement of the merger that consolidated the facilities and assets of the entire Mackay System with those of the International Telephone and Telegraph Corporation. Control of All American Cables, Inc., had been acquired in 1927. I.T.&T. had entered the manufacturing field in 1925 with the purchase of the International Western Electric Company, the name of which was subsequently changed to International Standard Electric Corporation, now the world-wide sales and manufacturing subsidiary of I.T.&T. Extension both of its telephone operating and manufacturing activities in Latin America, Europe, and Asia came in the ensuing years. Parallel developments took place in the field of international communication by cable and radio.

Palo Alto Plant Moved to Newark, New Jersey

In the summer of 1931, the Federal Telegraph Company was moved from Palo Alto to 200 Mount Pleasant Avenue, Newark, New Jersey, where it still continues many of its operations. During this period, its business consisted primarily in the manufacture of marine radio equipment for the Marine Division of Mackay Radio and Telegraph Company and the design and manufacture of high-power vacuum tubes, transmitters, and receivers used by Mackay Radio in its point-to-point communication system.

In 1932, a number of one-kilowatt, vacuum-tube radiotelegraph transmitters were designed and constructed for the China Electric Company. Although a pioneer in the construction of many earlier types of transmitters, this was the initial venture of any magnitude in the production of the modern type of transmitting equipment other than for marine use.

In 1934, a quantity of 50-kilowatt vacuum tube radiotelegraph transmitters was built for use by Mackay Radio in expanding its communication's network both within the U. S. A. and for communication with countries abroad. Apparatus was also manufactured for All America Cables and the Postal Telegraph Company, as well as transmitting tubes in increasing quantities.

In 1934, development of the first of a line of water-cooled tubes for use in broadcast services was completed. This marked the initial step in accepting vacuum tube business from non-associated companies and paved the way for Federal's present prominent position in the manufacture of tubes for most of the larger broadcasting stations throughout the United States.

Federal Begins Work on Defense Program

In 1936 the Maritime Commission's National Defense Program called for the construction of 500 ships. Federal developed a complete shipboard radio unit for installation in these vessels and has since manufactured the larger share of these assemblies which form a part of the standard equipment aboard Liberty and Victory cargo ships.[3] In 1937 Federal produced the prototype of the "walkie-talkie" for the U. S. Signal Corps, as well as considerable other equipment for the Government. From then on, growth was tremendous under the impact of requirements for defense and later for war purposes.

The International Telephone and Telegraph Corporation, having brought its aerial navigation and selenium rectifier developments to this country in 1938, organized the International Telephone Development Company the following year for their commercialization. In 1939, when hostilities broke out in Europe, the I.T.&T. Corporation found it necessary to establish a manufacturing unit in the United States for the purpose of supplying equipment to its telephone operating subsidiaries in the Western Hemisphere.

A factory was located in Newark, New Jersey, and the new organization, known as the International Telephone and Radio Manufacturing Corporation, absorbed the International Telephone Development Company, whose operations had been centered in New York. Into the I.T.-R.M.C. went outstanding developments long pioneered by I.T.&T. manufacturing associates in Europe. Belgian and French developments were brought to the U. S. A. just ahead of and, in some cases, even after Axis occupation.

The International Telephone and Radio Manufacturing Corporation acquired control of the Federal Telegraph Company in June, 1941, and the two operated as separate I.T.&T. associates until October, 1942, when they were merged into the Federal Telephone and Radio Corporation.

FTR Laboratories Formed

Coincidentally with these organizational moves, the Federal Telephone and Radio Laboratories were established as the research and development division of the Federal Telephone and Radio Corporation at 67 Broad Street, New York, N. Y. The nucleus of the Laboratories was a group of development engineers, specialists in the fields of immediate military needs, such as direction finders, aerial navigation, radar, and communication systems.

As the activities of the laboratories grew, they were allotted more and more space at 67 Broad Street and elsewhere. Field stations also were set up at Great River, Long Island; Telegraph Hill, Hazlett, New Jersey; and at the Rye Lake Airport, Westchester County, New York, for flight-testing airborne equipment.

World War II Developments

A few months before United States entry into World War II, a new 50-kilowatt main transmitter and a 5-kilowatt emergency transmitter for WABC, key station of the Columbia Broadcasting System, and now WCBS, were completed.[4] The new station was opened October 18, 1941, and was described as the last word in modern high-fidelity amplitude-modulation broadcasting equipment. From December 7, 1941 to the end of hostilities, it operated on a continuous 24-hour schedule. In January, 1942, the Columbia Broadcasting System also inaugurated its new international short-wave broadcasters built by Federal.[5, 6]

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[3] E. J. Girard, "A New Marine Radio Unit for Cargo Vessels," *Electrical Communication*, v. 20, n. 2, pp. 71–72; 1941. See also: E. J. Girard, "The H. F. Marine Radio Unit—Newly Adopted Equipment Influencing Ship Design," *Electrical Communication*, v. 21, n. 2, pp. 85–88; 1943. See also page 468, this issue of *Electrical Communication*.

[4] E. M. Ostlund, "WABC—Key Station of the Columbia Broadcasting System," *Electrical Communication*, v. 21, n. 1, pp. 61–71; 1942.

The station consisted of three 50-kilowatt final amplifiers and two 50-kilowatt modulators arranged for simultaneous operation and automatic switching of frequencies to 13 directive arrays capable of operating between 6 and 22 megacycles. Later, a third 50-kilowatt modulator was completed so that three programs could be broadcast at one time, beamed to South America, Central America, the West Indies, Mexico, Europe, Africa, and Asia.

During the war, this station was operated around the clock by the Office of War Information and programs were broadcast in 23 languages.

Development of a high-power, high-frequency vacuum tube for short-wave broadcasting made possible the construction of 200-kilowatt stations.[7, 8] Federal built two of these 200-kilowatt short-wave broadcast transmitters.[9] They were

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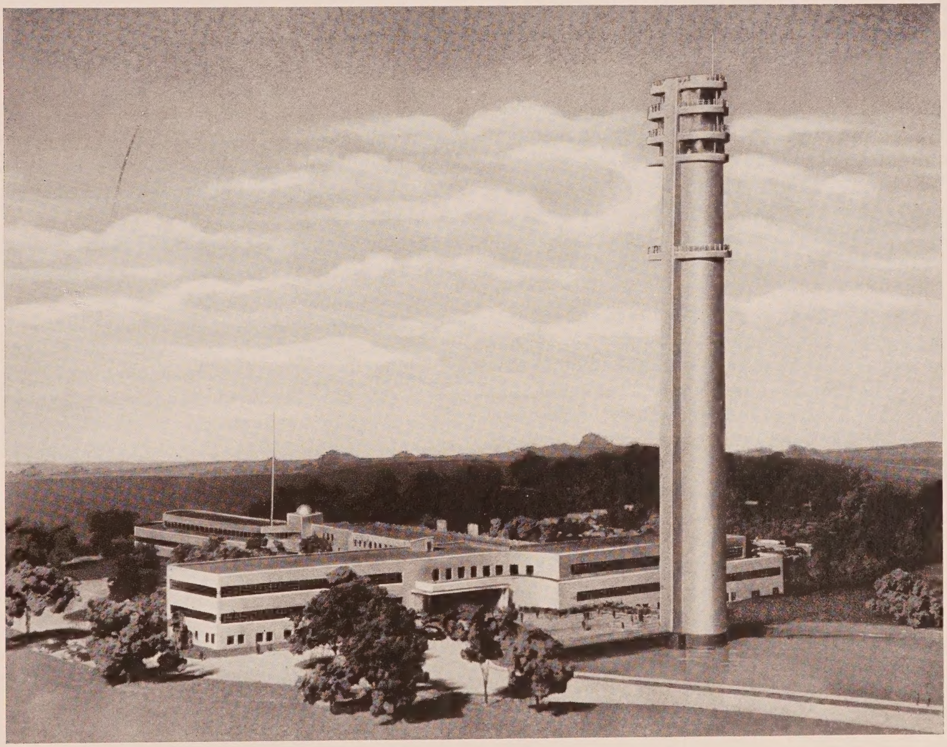
[5] H. Romander, "New 50-kilowatt CBS International Broadcasters," *Electrical Communication*, v. 21, n. 2, pp. 112–123; 1943.

[6] Howard A. Chinn, "Audio and Measuring Facilities for the CBS International Broadcast Stations," *Electrical Communication*, v. 21, n. 3, pp. 174–179; 1943.

[7] G. Chevigny, "Tubes for High-Power Short-Wave Broadcast Stations," *Electrical Communication*, v. 21, n. 3, pp. 143–156; 1943.

[8] "A Proposal for a Global Shortwave Broadcasting System," *Electrical Communication*, v. 22, n. 2, pp. 154–166; 1944.

[9] H. Romander, "200-Kilowatt High-Frequency Broadcast Transmitters," *Electrical Communication*, v. 22, n. 4, pp. 253–261; 1945.



Architect's model showing the new buildings and microwave tower for Federal Telecommunication Laboratories, scheduled to be completed in 1947. The first unit was opened in October, 1945.

placed in operation on the west coast in 1945 and performed an important function in the beaming of programs to the Pacific area and the Orient.

In 1942, manufacture was started of a new improved type of subscribers' desk set employing plug-in tropicalized components.[10] This desk set assumed an important war role when, in 1943, it was specified as standard equipment for tropical installations by the United States Signal Corps.

Already operating as an "Arsenal of Democracy" before Pearl Harbor, Federal soon was working full time on war equipment. It took the lead in the design of high-power pulse radar triodes.[11]

Classified as one of the most important developments of the war and, incidently, one of its best kept secrets was Federal's high-frequency instantaneous direction finder. Credited by the Navy as an important factor in breaking up the submarine wolf-pack warfare, this "secret weapon" was affectionately labelled "Huff-Duff."

Federal's role in the production of telephone equipment for the Armed Forces, while not dramatic, was large and varied. It ranked first in the production of field telephone sets and provided substantial quantities of switchboards, power plants, and repeaters. In many cases Federal redesigned the equipment for simplification, lighter weight, tropicalization, or wider application.

Its production of selenium rectifiers for war equipment was enormous. The variety of uses as well as their efficiency were stepped up so that selenium rectifiers were important equipment in almost every type of war implement. Significant among these were selenium rectifiers developed for the proximity fuse.

It also was the first company to produce a solid-dielectric high-frequency transmission line, meeting all requirements of the Armed Services. It made the only commercially successful dual coaxial transmission line used for direction finding and instrument landing systems and it developed, in conjunction with the Radiation Laboratory at Massachusetts Institute of Technology, the first spiral delay line. It pioneered in the development of low-capacitance transmission cable.[12, 13]

The story of Federal's wartime activities, or at least the part that no longer remains classified, has been outlined in this journal.[14, 15] The Army and Navy "E" was awarded the Federal Telephone and Radio Laboratories on April 20, 1943. A second star was awarded December 2, 1944 and a third—April 11, 1945. The Federal Telephone and Radio Corporation received the Maritime "M" Pennant and Victory Fleet Flag, December 21, 1942, and the Army and Navy "E" award, January 30, 1945.

Plans for Modern Buildings

Because of the corporation's exceptional growth, centralization of manufacturing facilities during World War II was impossible, and activities were spread over 50 locations, including plants, storage accommodations, and administrative quarters, located in Newark, New Jersey, and adjacent areas. Nevertheless, plans for the construction of modern buildings to house the manufacturing activities as well as laboratories were formulated. The 120-acre site of the former Yountakah Country Club in Nutley and Clifton, New Jersey, was purchased for this purpose. Ground for the first unit was broken on February 17, 1943.

The new laboratories are located at the Nutley area of the site, and the manufacturing plant in Clifton. First units of both groups of buildings were formally dedicated October 3, 1945. The laboratories, embodying advanced construction theories of architects and scientists, are of steel fabrication sheathed with aluminum, a type of design which affords protection from outside electrical disturbances. When these laboratory buildings are completed early in 1947, all Federal research activities will be centered in Nutley.

A special 300-foot tower is also being erected at Nutley for ultra-high-frequency research on radar, aerial navigation, frequency modulation, pulse-time modulation, television, mobile radio, and point-to-point long-distance telephony.

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[10] E. S. McLarn, "Simplified Subscribers' Telephone Sets," *Electrical Communication,* v. 21, n. 1, pp. 3–12; 1942.

[11] J. J. Glauber, "Radar Vacuum Tube Development," *Electrical Communication,* v. 23, pp. 306–326; September, 1946.

[12] H. Busignies, "Applications of H-F Solid-Dielectric Flexible Lines to Radio Equipment," *Electrical Communication,* v. 22, n. 4, pp. 295–301; 1945.

[13] K. H. Zimmermann, "Spiral Delay Lines," *Electrical Communication,* v. 23, pp. 327–328; September, 1946.

[14] *Electrical Communication:* 1940–1945—Part II, v. 23, pp. 214–240; June, 1946.

[15] *Electrical Communication:* 1940–1945—Part III, v. 23, pp. 363–366; September, 1946.

The manufacturing plant at Clifton when completed will provide approximately one million square feet of floor space.

Post-War Developments

Soon after V-J day (on September 27, 1945), Federal demonstrated a new system of Pulse-Time Modulation. For the first time, using PTM, 24 simultaneous 2-way conversations were carried over an 80-mile microwave relay circuit at 1300 megacycles.[16] All conversations were transmitted on a single carrier frequency. An important feature of PTM is that only one transmitter and receiver are employed at each terminal for all conversations. Further, a PTM demonstration in 1946 showed the feasibility of broadcasting as many as eight programs of high-fidelity speech and music, facsimile, and teleprinter signals, from a single transmitter.[17] A single receiver detected, amplified, and separated the eight programs for distribution to the loudspeakers and other terminal equipment.

In January, 1946, Federal installed the first high-definition, ultra-high-frequency, full-color-television transmitter in the Chrysler Building, New York City, for the Columbia Broadcasting System. The new transmitter operates on 490 megacycles with one kilowatt of power.[18]

In October, 1946, Federal demonstrated features of its new aerial navigation and traffic control system.[19] The new system embodies the results of a concentrated effort in the direction of coordinated planning, global in scope, of radio aids to aerial navigation. It supplements Federal's pioneer work in developing instrument landing equipment for the Civil Aeronautics Authority and the United States Signal Corps, as well as its development of the ultra-high-frequency radio range with sector identification.[20, 21, 22]

Along with its early work in developing A-M broadcasting transmitters, Federal is now taking a leading position in the production of F-M broadcast transmitters. Federal also has developed a new 8-element square loop antenna for F-M stations which gives an effective power gain[23] of 9.

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14. H. Pratt, "Long Range Reception with Combined Crystal Detector and Audion Amplifier," *Proceedings of the I.R.E.*, v. 3, pp. 173–178; June, 1915.
15. "Report of the Federal Trade Commission on the Radio Industry," Government Printing Office, Washington, D. C.; 1924.

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[16] D. D. Grieg and A. M. Levine, "Pulse-Time-Modulated Radio Relay System—Terminal Equipment," *Electrical Communication*, v. 23, pp. 159–178; June, 1946.

[17] D. D. Grieg, "Multiplex Broadcasting," *Electrical Communication*, v. 23, pp. 19–26; March, 1946.

[18] See Page 406, this issue of *Electrical Communication*.

[19] H. Busignies, Paul R. Adams, Robert I. Colin, "Aerial Navigation and Traffic Control with Navaglobe, Navar, Navaglide, and Navascreen," *Electrical Communication*, v. 23, pp. 113–143; June, 1946.

[20] H. H. Buttner and A. G. Kandoian, "Development of Aircraft Instrument Landing Systems," *Electrical Communication*, v. 22, n. 3, pp. 179–192; 1945.

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[22] A. Alford, A. G. Kandoian, F. J. Lundburg, and C. B. Watts, Jr., "An Ultra-High-Frequency Radio Range with Sector Identification and Simultaneous Voice," *Electrical Communication*, v. 23, pp. 179–189; June, 1946.

[23] See page 415, this issue of *Electrical Communication*.

Color-Television Transmitter for 490 Megacycles

By N. H. YOUNG

Federal Telecommunication Laboratories, Inc., New York, New York

A TELEVISION transmitter, developed to specifications of the Columbia Broadcasting System, is described. It radiates color-television signals at a carrier frequency of 490 megacycles per second with a power output of approximately 1 kilowatt.

• • •

1 *Requirements*

The image analysis used by the Columbia Broadcasting System requires the transmission of all frequencies between 20 cycles per second and 10 megacycles for faithful reproduction of the picture. The carrier frequency of 490 megacycles was assigned by the Federal Communications Commission. The power output was to be the highest that could be generated reliably with the tubes and techniques immediately available. Allowing for the concentration of power possible with directive antennas on the ultra-high-frequency band, it was believed that a power of the magnitude of 1 kilowatt would be adequate for satisfactory coverage of the New York metropolitan and suburban areas. This has been confirmed by field tests that are still in progress. In view of the severe electrical requirements and because of the adequate facilities available on the 71st floor of the Chrysler Building, no rigid space or weight limitations were imposed. The specifications for the transmitter are given in

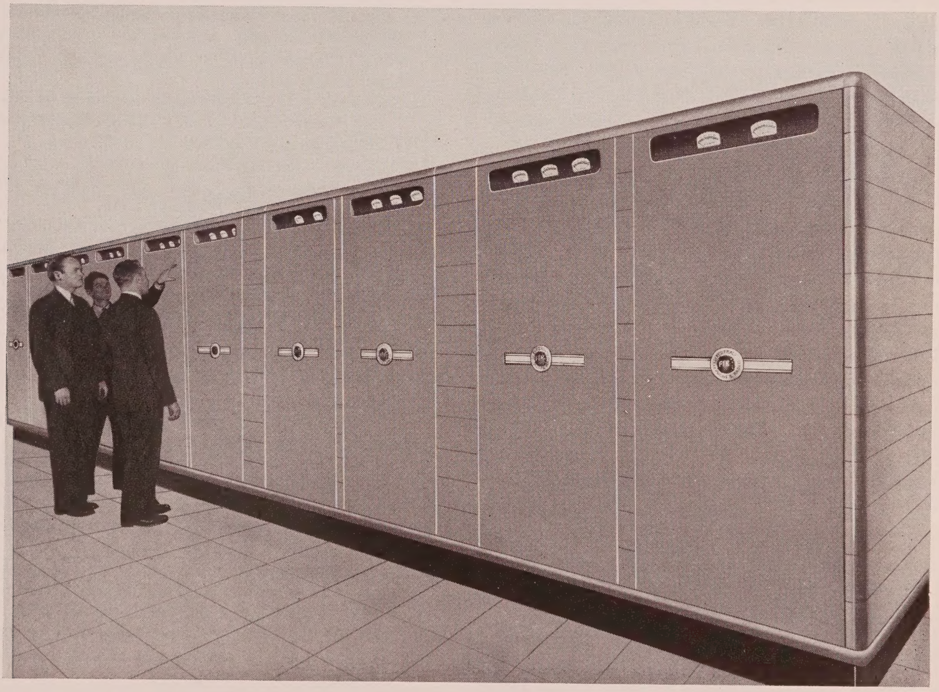

Fig. 1—Federal and Columbia Broadcasting System engineers inspect the completed transmitter as installed on the 71st floor of the Chrysler Building in New York City.

Table I. Before discussing the design details of the individual units of the transmitter, general information about the transmitter and its components may be obtained from the picture of the installation at the Chrysler Building, Fig. 1, and from the block diagram shown in Fig. 2.

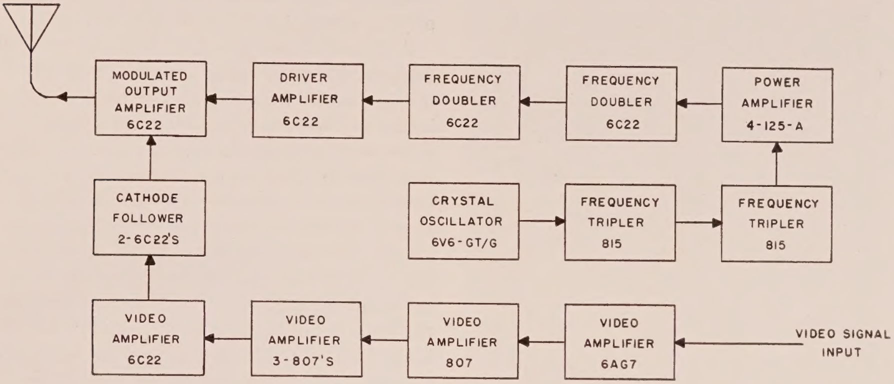
MODULATED OUTPUT AMPLIFIER 6C22
DRIVER AMPLIFIER 6C22
FREQUENCY DOUBLER 6C22
FREQUENCY DOUBLER 6C22
POWER AMPLIFIER 4-125-A
CATHODE FOLLOWER 2-6C22'S
CRYSTAL OSCILLATOR 6V6-GT/G
FREQUENCY TRIPLER 815
FREQUENCY TRIPLER 815
VIDEO AMPLIFIER 6C22
VIDEO AMPLIFIER 3-807'S
VIDEO AMPLIFIER 807
VIDEO AMPLIFIER 6AG7
VIDEO SIGNAL INPUT

Fig. 2—Block diagram of color-television transmitter.

2 *Output Stage*

TABLE I

SPECIFICATIONS FOR THE TRANSMITTER

Carrier Radio Frequency (megacycles)	490
Peak Power Output (watts)	1,000
Maximum Average Power Output (black-level condition, watts)	600
Frequency Stability (percent)	±0.005
Effective Modulation Bandwidth, 30 cycles to 10 megacycles (decibels)	+1 to −3
Phase Response. Maximum delay difference between any two frequencies between 30 kilocycles and 10 megacycles (microseconds)	0.05
Video-Frequency Input Required (volts, peak-to-peak, white positive)	2
Video-Frequency Input Impedance (ohms)	70, adjustable ±5 ohms to match exactly the video-frequency input line

The most critical point in the design of a television transmitter is the choice of tube, circuit, and operating conditions for the output radio-frequency amplifier as this will determine the maximum power output. The most important factors are the bandwidth to be transmitted, the carrier frequency, and the output power. In most cases this power is limited by the emission of the tubes in the output stage and more detailed information on this phase of design will be found in a previous paper.[1] In this paper the following relation is set forth.

$$I_s = 0.4\sqrt{PCf}, \qquad (1)$$

where

I_s = saturation current in amperes,
P = peak power output in kilowatts,
C = total output circuit capacitance in micromicrofarads,
f = maximum modulation frequency in megacycles.

Applying this formula to the 6C22 tube for color-television conditions,

$I_s = 5$ amperes,
$C = 6.4$ micromicrofarads,
$f = 10$ megacycles,

hence

$$5.0 = 0.4\sqrt{P \times 6.4 \times 10},$$

and

$$P = 2.4 \text{ kilowatts}.$$

This shows that, viewed from the standpoint of emission, capacitance, and bandwidth, the

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[1] E. Labin, "Design of the Output Stage of a High Power Television Transmitter," *Electrical Communication* , v. 20, n. 3, pp. 193–201; 1942.

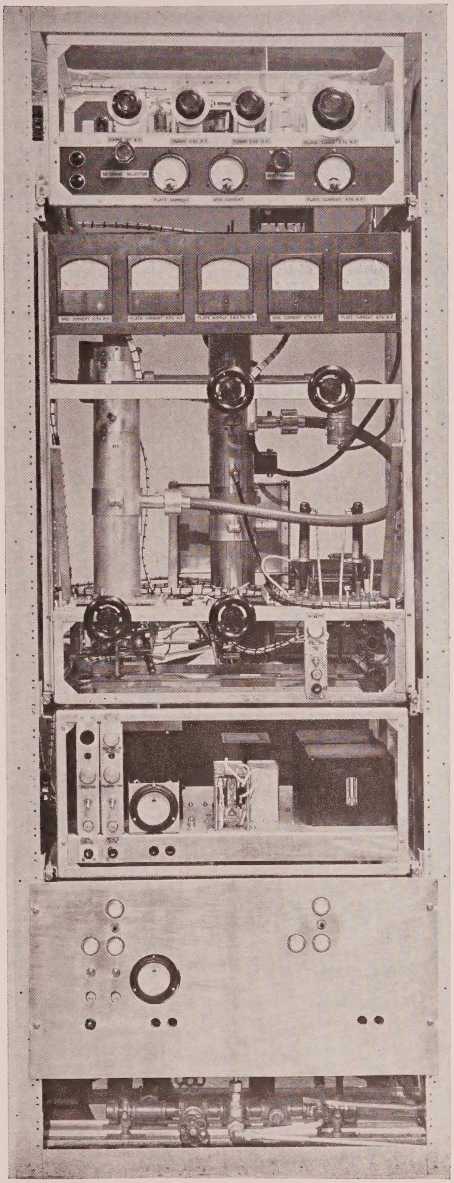

Fig. 3—A single cabinet accommodates the 6.805-megacycle crystal oscillator and following stages, which multiply frequency and power to 490 megacycles and 300 watts to drive the penultimate and final stages.

tube should be capable of generating 2450 watts peak output. This analysis, however, does not include the dissipation capabilities of the tube. In this case, the limiting factor is the dissipation of energy in the control grid of the tube. When this limitation is included, the peak power output possible is found to be 1200 watts. The final transmitter is conservatively rated at 1000 watts of peak power, insuring full tube life.

A second point of major importance in the design of television transmitters is the bandwidth of the output circuits. This has been considered in the equation above, insofar as it affects the power output capabilities of the tube, but attention must be given to the methods by which the circuit may be adjusted to a suitable Q. The design of the 6C22 is such that normal operation is obtained at relatively low plate voltages and high currents; the tube plate resistance is low. Thus an appreciable portion of the damping required in the output tank circuit is supplied by the tube itself. In addition, the coupling of the load circuit for optimum output results in a lower coupled impedance to match the low plate resistance of the tube, so that the load damps the output circuit more heavily than would be the case with other tube designs. The combination of these effects is sufficient to permit the operation of the output stage of the color-television transmitter with no damping other than that provided by the tube and the load. Attenuation of sidebands in the output stage is less than 2 decibels at 10 megacycles from the carrier.

3 *Radio-Frequency Exciter*

For fixed stations, where considerations of power supply, space, and weight are not extremely severe, and for operation on one frequency only, there is no doubt that the most reliable and accurate control of frequency is obtained with crystal control of the fundamental oscillator. This is used in the transmitter.

The tube lineup was determined by working backward from the output stage, based on the excitation requirements of each stage in turn and on the types of tubes commercially available for each position. While in some cases a number of solutions were possible, it is believed that the one chosen represents a reasonable compromise.

The oscillator stage uses a 6V6GT/G tube in a tri-tet circuit. The crystal is operated at 6.805 megacycles and the frequency is doubled in the plate circuit of the same stage, resulting in an output of approximately 3 watts at 13.611 megacycles.

The following stage is a push-pull frequency tripler using a type 815 tube. The 815, a dual beam tetrode, delivers approximately 10 watts at 40.833 megacycles. The next stage is a similar tripler using the same type of tube and delivers approximately 10 watts at 122.5 megacycles.

The fourth stage is a conventional amplifier using a type 4-125 power tetrode. This tube is capable of operating as an amplifier at frequencies up to 125 megacycles without neutralization, although it is necessary in this case to resonate the screen-cathode connection to bring the screen effectively to ground potential and minimize the effects of lead inductance. This arrangement has proved to be entirely stable and the stage delivers approximately 120 watts at 122.5 megacycles.

The four stages described are mounted on one chassis and designated as the exciter unit of the transmitter. This unit and its power supply are located in the first radio-frequency cabinet of the final assembly.

The remaining stages of the radio-frequency amplifier chain use the 6C22 tube.[2] This is the only tube currently available for powers of 200 to 800 watts at frequencies between 300 and 600 megacycles. The tube, which is a triode of high mutual conductance and low plate resistance, uses the ring-seal technique to insure satisfactory operation at ultra-high frequencies.

The fifth stage of the radio-frequency chain consists of a 6C22 in a coaxial circuit operating at a high power level. The cathode of the tube is by-passed to ground as directly as possible and excitation is furnished to the grid. The tuning circuit for the grid is a section of coaxial line fitted with a piston and a capacitive tuning slug. By proper adjustment of the piston and slug, it is possible to have the grid at a high impedance above ground at the input frequency, and at the same time at a very low impedance to ground at the output frequency. This eliminates all tendency to instability in the stage. When excited from the unit previously described, the stage delivers 250 watts output at 245 megacycles.

The sixth stage is also a frequency doubler using a 6C22. Because of lead inductance, it is no longer possible to ground the cathode at this higher frequency. Rather than extend the filament line by a half wavelength, making a rather large and clumsy assembly, a frequency doubler was designed for grounded-grid operation. Here the excitation is applied to the cathode, the grid is grounded, and power is taken from the plate circuit. This stage delivers approximately 300 watts at 490 megacycles. A view of the first six stages mounted in a rack is shown in Fig. 3.

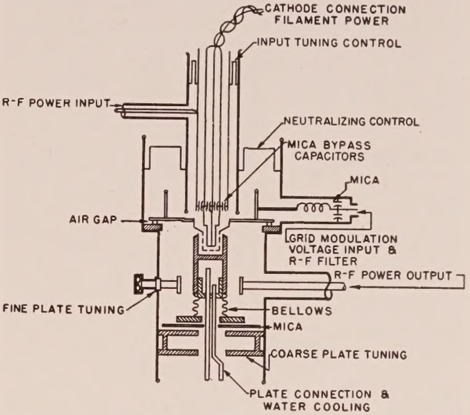

Fig. 4—Sectional view of neutralized power amplifier for 490 megacycles as used in seventh and eighth radio-frequency stages.

The seventh stage is a neutralized power amplifier using the 6C22 in a coaxial grounded-grid circuit. With 300 watts of drive it delivers approximately 700 watts of energy at 490 megacycles. This is considerably more than would be required to drive the final modulated stage to full output, but the excess power is dissipated in a damping resistor attached to the coupling line between the driving amplifier and the output stage. The load imposed by this resistor acts to maintain a constant output voltage from the driving stage notwithstanding the changing load of the modulated stage as its grid bias is varied through the modulation cycle. This improves the linearity of the modulation characteristic and somewhat reduces the magnitude of the voltage required from the modulator.

The eighth stage is the grid-modulated output amplifier, the design of which has been discussed already. It delivers a modulated television signal of 1 kilowatt peak, and up to 600 watts average, depending on signal content, with a modulation bandwidth of ±10 megacycles. A sectional diagrammatic view of the amplifier used in the seventh and eighth stages of the transmitter and the assembly are shown in Figs. 4 and 5. Some idea of the complexity of the anode and grid circuits in this assembly may be gained from the "exploded" view in Fig. 6.

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[2] J. J. Glauber, "Radar Vacuum-Tube Developments," *Electrical Communication*, v. 23, pp. 308–321; September, 1946.

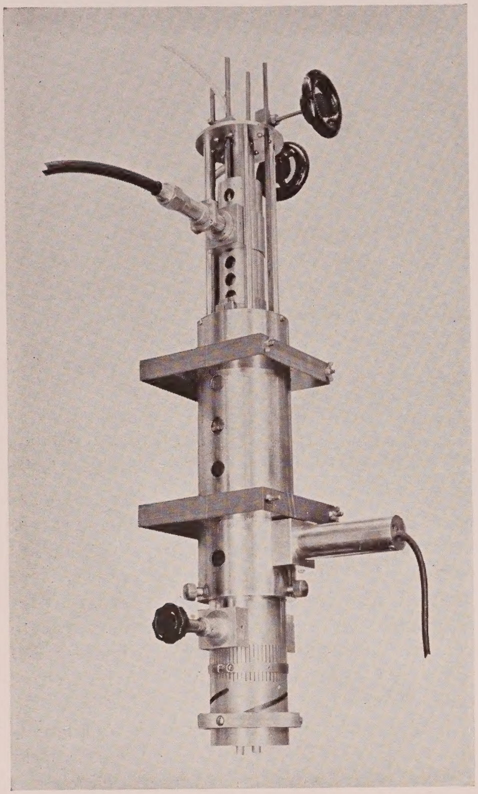

Fig. 5—Coaxial amplifier using 6C22 tube at 490 megacycles.

4 Modulator

In the design of modulation systems for television transmitters, two principal problems are encountered which may be solved almost independently. Basically these are the response of the amplifier to very high frequencies and to very low frequencies.

In meeting the requirements for delivering energy at very high frequencies, the problem resolves itself into that of providing a specified voltage across a small impedance. The impedance cannot be increased because of inevitable capacitive shunting of essential circuit elements, so rather large current swings are required to provide this voltage. The problem is naturally most difficult in the output stage, because this is where the highest voltage is required, and usually also the point where the capacitances are least under control of the designer.

In the output stage of this modulator, this problem has been met by the use of a cathode follower, employing two 6C22 tubes in parallel. Because of the negative feedback present in this circuit, its gain is substantially uniform throughout the video-frequency band in spite of the relatively large capacitance of the modulated-stage grids and the connecting cable. The provision of two tubes in parallel enables the stage to furnish

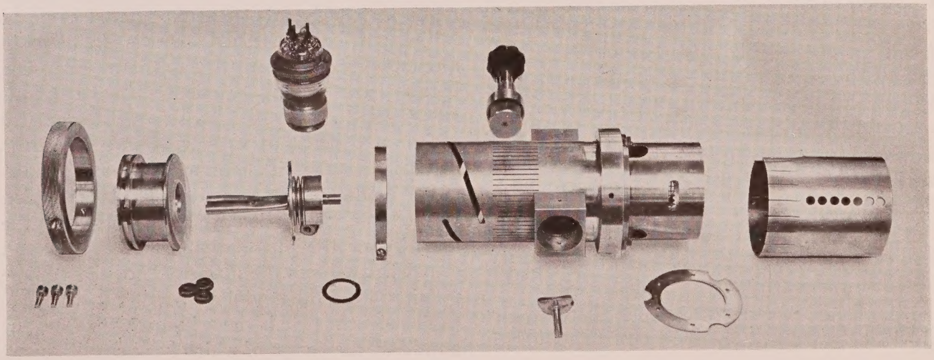

Fig. 6—Exploded view of coaxial amplifier using 6C22 tube at 490 megacycles.

the required current without danger of overload at the higher modulation frequencies.

In the stages feeding the modulator, the high-frequency response and gain have been maintained as well as possible by designing the interstage coupling networks according to the principles of low-pass filter theory.[3] The use of four-terminal networks in some locations enables the splitting of the output capacitance of one stage from the input capacitance of the following stage permitting a higher impedance load to be used without degrading the high-frequency response.

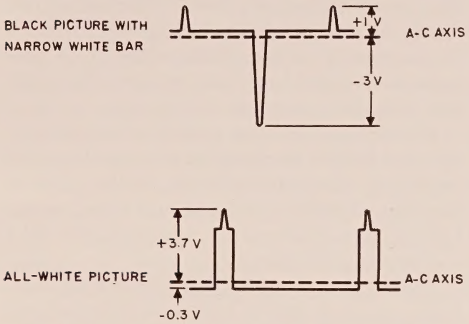

Fig. 7—Comparison of direct- and alternating-current coupling applied to typical television waveforms.

The amplifier is also unusual in its low-frequency coupling methods. The varied waveforms which may be present in television image signals, make it possible for the alternating-current axis of the wave to lie near either the positive or negative peak of the wave, or at any position in between. For example, let us consider Fig. 7, which shows extreme values which normal television waveforms might assume. Considering a video-frequency signal of 4 volts peak to peak, with the usual ratio of 25-percent synchronizing pulses, the dimensions would be as shown in the drawing. In conventional amplifiers using capacitive coupling, the alternating-current axis is maintained at a fixed point on the tube characteristic for all waveforms, and to accommodate the signals shown, a stage would have to accept without distortion all potentials between $+3.7$ and -3 volts from that point, a total swing of 6.7 volts.

On the other hand, if the amplifier is coupled for the direct-voltage components of the signal, or if a "d-c restorer" circuit is used, the peaks of the synchronizing signal will always be brought to the same point of the tube characteristic. Regardless of image content, the potential must lie in a negative direction within 4 volts of this value. Thus it will be seen that the following stage will be required to accept inputs over a range of only 4 volts. The inclusion of the direct-voltage component results in a saving in swing requirements of more than 40 percent.

Since the provision of adequate swing capability is one of the most difficult phases of the amplifier problem, particularly in high-powered stages, it is economically essential that the direct current be present in all the later stages of the modulator. If direct-current restoring circuits are used, one must be provided for each stage. Unfortunately, these circuits embody time constants and the cascading of their effects might tend to produce undesired slopes on portions of the signal wave. For this reason it was decided to retain the direct-current component through the interstage coupling networks, thus eliminating any need for restoring it later.

4.1 Coupling Pack

The method of low-frequency coupling is shown in Fig. 8. For simplicity, no high-frequency compensation is shown in the figure, but it may be added without affecting the result. A capacitor $C1$ is connected from the plate of the first stage to the grid of the second. From each side of this

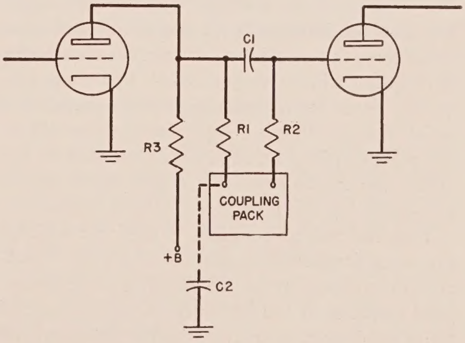

Fig. 8—Method of coupling for low frequencies.

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[3] H. A. Wheeler, "Wide-Band Amplifiers for Television," *Proceedings of the I.R.E.*, v. 27, pp. 429–437; July, 1939.

capacitor, isolating resistors, $R1$ and $R2$, are connected to the terminals of a regulated power supply, referred to as a coupling pack. No other path is provided from the grid of the second stage to ground. The cathode of the second stage is directly grounded. The isolating resistors are much larger than the first-stage load resistor $R3$ so that at the very high frequencies, where the capacitance to ground of the regulated pack represents a very low impedance, the total shunting effect of $R1$ and $R2$ in parallel with $R3$ will not be very much lower than $R3$ alone. Thus, for high frequencies, the terminals of the coupling pack will remain at constant potential and the signal will be transferred to the grid of the second stage through the coupling capacitor $C1$.

For very low frequencies, the impedance of $C1$ will be high in comparison with $R2$ so that coupling through the capacitor alone will have little effect. However, at low frequencies the capacitance to ground of the coupling pack will represent a high impedance in comparison with $R1$, so that the positive terminal of the coupling pack will assume the potential of the plate of the first stage. Thus for very low frequencies, the signal is coupled from the plate of the first stage to the grid of the second through a path consisting of $R1$, the coupling pack, and $R2$, with both terminals of the coupling pack fluctuating above ground in accordance with the signal energy.

As the values of $R1$ and $C2$ enter into the equation for the frequency at which the coupling pack ceases to follow the fluctuations of the signal, and the values of $R2$ and $C1$ determine the frequency where the coupling through the capacitor ceases to be effective, it is only necessary that the product of $R1$ and $C2$ shall be very much less than the product of $R2$ and $C1$. Under these conditions, the transition between the two modes of coupling will be very gradual. For every frequency from zero to the highest frequency for which the amplifier is compensated, there will be adequate coupling through either one path or the other, and in the transition region there will be a re-enforcing coupling through both paths.

This method of coupling is used throughout the video-frequency system to retain the direct-current component of the signal and to insure good response at the low frequencies. In spite of the possibility of a small instability in one of the early stages being magnified to cause "bounce" or flicker in the output, no such effects have been observed, principally because of thorough attention to regulation of the coupling packs and plate supplies in the low-level stages of the system. The use of shielded transformers in the coupling packs is required to prevent the introduction of "hum" frequencies into the system.

4.2 Modulator Stages

The first stage of the modulator uses a 6AG7 tube. Normal input is 2 volts. (Throughout this paper, all video-frequency voltages are referred to in terms of peak-to-peak values. This conforms to the usual television practice.) The stage gain is 7, giving an output of 14 volts.

The second stage uses an 807 tube, providing a gain of 2.8 and an output of 40 volts. The third stage uses three 807 tubes in parallel. This is necessitated by the relatively large input capacitance of the following stage. The stage gain is 4.5 and the output is 180 volts.

The fourth stage uses a 6C22 as a conventional triode amplifier. Although the tube interelectrode capacitances are not large, the Miller effect greatly increases the apparent input capacitance of the stage. With a suitable driving stage, no other detrimental effects are found. The gain of the stage is 3.5 and the output is 700 volts.

The fifth stage is a cathode follower using two 6C22 tubes in parallel. The principal function of this stage is to supply a signal from a source of sufficiently low impedance so that the effects of changing load imposed by the output-stage grid current will be negligible. In the region where they are operated, each tube has a mutual conductance of 10 000 micromhos, and the source impedance of this stage may be considered as 50 ohms. In addition, the high current capabilities of the stage and the negative feedback present in the cathode-follower connection provide a flat response-frequency characteristic in spite of the shunt capacitance of the radio-frequency amplifier load. The stage gain is 0.8 and the output voltage is 550 volts.

5 *Power Supply and Control Problems*

In any experimental transmitter, one of the most important features of the control system must be a flexibility sufficient to permit all normal and abnormal operations that might be required in tests or experiments. On the other hand, for daily service in broadcasting, it is desirable that only a few controls be provided, as this makes for trouble-free operation and prevents errors in the starting procedure.

These two requirements have been combined in this transmitter. The power control panel is shown in Fig. 9. In a normal installation, all the controls except the 5 toggle switches across the top may be covered by a door, and need not be touched in turning the transmitter on and off. When the main power switch and the pump have been turned on, operation of the switch marked "Excitation" results in the application of power to all the stages except the final radio-frequency amplifier and the final stage in the modulator. Where necessary, time-delay relays and interlocking relays are provided to protect the equipment and maintain the proper operating procedure.

For experimental work or testing, however, by opening the door exposing the lower panel of controls and pilot lights, each unit may be controlled individually. The automatic starting procedure may be used up to any desired point, at which it may be arrested by opening one or more of the switches. In all cases, interlocks prevent incorrect operation and protect equipment and personnel.

In the design of the power supplies, conventional practice was followed. The plate supplies for the first three video-frequency amplifiers and all the coupling packs used in the direct-coupled amplifier are electronically regulated to stabilize output voltage against changes in load or supply voltage. All other supplies use simple rectifiers and low-pass filters of conventional type, although the output capacitance of certain of the filters is much larger than would be required in the design of a transmitter for audio frequencies.

All plate-supply leads carrying video-frequency currents are run in coaxial cable of 50-ohm characteristic impedance and terminated at the *power-supply* end in a resistive load. Attempts to use simple wire lines with by-pass capacitors at the amplifier end resulted in resonance at some point in the video-frequency range, producing undesirable irregularities in the response-versus-video-frequency curve. This arrangement was required in all stages of the modulator, and in the plate and bias leads of the modulated radio-frequency amplifier stage. Where several stages are operated from the same power supply, individual cables and terminations are required for each.

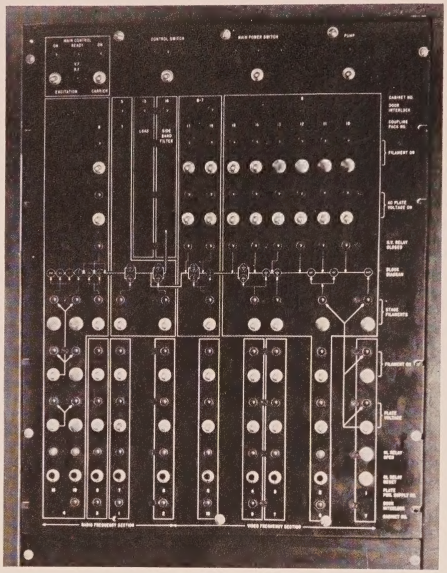

Fig. 9—Power control panel for television transmitter.

An additional desirable effect is obtained by operating the fourth and fifth stages of the modulating amplifier from a common plate supply. As the current drawn by one stage is increasing when that drawn by the other is decreasing, a compensating effect is obtained, making the problem of power-supply filtering easier for very low frequencies.

6 Facilities for Installation

The complete transmitter is contained in six standard transmitter bays, plus a control console. All power supplies are included in this space, but a pump and heat exchanger for the water cooling system must be provided elsewhere.

The equipment is designed for operation from either 208 or 220 volts, 3 phase, 60 cycles. A total of 25 kilovolt-amperes is drawn and includes the auxiliaries mentioned above.

7 Conclusion

In completing the construction of a transmitter meeting the exacting requirements of color television in the ultra-high-frequency range, many problems of television technique and ultra-high-frequency construction have been solved. Based on this experience and on the material compiled in the laboratory testing of the finished transmitter, it will be possible to design other transmitters of similar characteristics as needed for all wide-band applications, whether for television, for multiplex pulse operation, for experimental broadcasting, or for military applications.

In addition, it should be noted that where transmitters of this type are used, the large bandwidth available makes it possible to transmit the sound associated with the television program through the same transmitter. This is accomplished by adding suitable multiplexing equipment in the studio. No alteration in the transmitter is required. This multiplexing may utilize pulse-time modulation, and field tests using this method have been made. Work is now in progress to determine the relative merits of pulse-time modulation and other multiplexing systems for this service.

Square Loops for Frequency-Modulated Broadcasting at 88–108 Megacycles

By R. F. LEWIS

Federal Telephone and Radio Corporation, Newark, New Jersey

FREQUENCY-MODULATED broadcasting in the U.S.A. has recently been assigned frequencies between 88 and 108 megacycles per second. Antennas for this band may be of relatively small physical size, and important transmission power gains may be obtained by designing antennas to give suitable directive radiation patterns.

This power gain can be obtained relatively easily as a comparatively large number of driven elements may be stacked on a single vertical mast. Such power gain must be controlled to increase the horizontal field pattern while reducing radiation in the vertical direction. For broadcast service to a limited area (primary service to a radius of 50 miles), where skywave transmission is not a factor, there are certain essentials that must be considered.

A. Circular pattern in the horizontal plane,
B. Good over-all impedance characteristics,
C. Maximum bandwidth consistent with cost,
D. Minimum number of selective circuits,
E. Possibility of stacking elements to increase gain,
F. Transmission-line system,
G. Ease of installation,
H. Minimum of maintenance, and
I. Low cost.

It is the purpose of this paper to consider the above features as pertaining to the square loop and the square-loop array. Items *A* to *E* are considered electrical problems. Item *F* includes both electrical and mechanical problems; the remaining items cover mechanical considerations alone.

• • •

The use of loop radiators for frequency-modulated broadcasting stems logically from earlier work having somewhat specialized applications.[1] Further developments, covered in a more recent paper,[2] have indicated the increased significance of these antennas particularly for producing circular radiation patterns.

1 Electrical

1.1 Circular Horizontal Pattern

An essentially circular pattern in the horizontal plane may be obtained from loops having small diameter with respect to wavelength. Small loops, however, present a very difficult problem from an impedance standpoint, especially if the bandwidth considered is even a few percent of the transmission frequency.

In Fig. 1 is a loop having straight sides, which give rise to the descriptive term "square loop." The dotted arrows represent the phase relations of the currents in each of the four sides. The dotted lines represent the current distribution along the radiating elements. The dimension of each side is a half wavelength, and the currents in each of the four sides are in phase. Such a configuration will give an essentially circular radiation pattern. Measurements have shown that

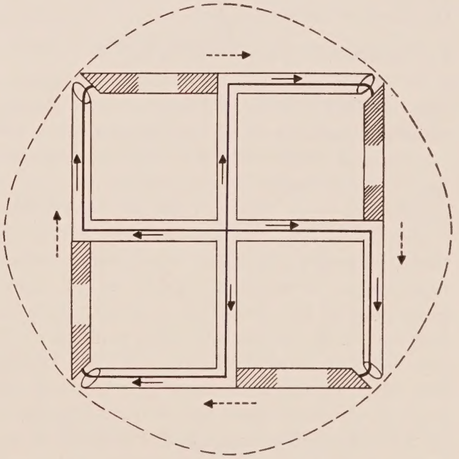

Fig. 1—Square loop, showing arrangements of coaxial transmission lines. Current phase and distribution are indicated by dotted arrows and lines.

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[1] A. Alford and A. G. Kandoian, "Ultra-High Frequency Loop Antennas," *Transactions American Institute Electrical Engineers*, v. 59, pp. 843–848; 1940: and *Electrical Communication*, v. 18, pp. 255–265; April, 1940.

[2] A. G. Kandoian, "Three New Antenna Types and Their Applications," *Proceedings of the I.R.E. and Waves and Electrons*, v. 1, pp. 70–75; February, 1946: and *Electrical Communication*, v. 23, pp. 27–34; March, 1946.

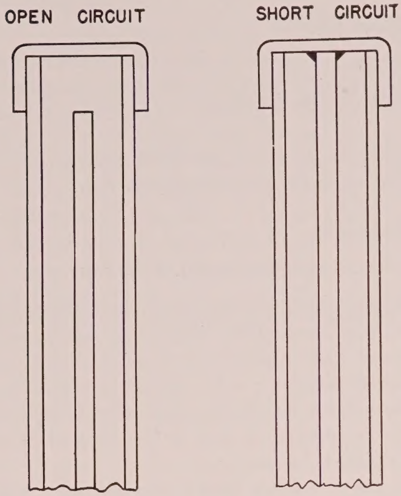

Fig. 2—Coaxial stub sections used for reactance matching. Both open- and short-circuited types are necessary to cover the 88–108-megacycle band adequately.

circularity in the horizontal plane can be maintained to within 1 decibel.

To excite this loop and obtain the current distribution desired, the coaxial transmission lines depicted in the figure are used. The feed point of each of the four transmission lines is looking at a point of high impedance. At the center of the square, the four lines are connected in parallel to the transmitter line.

1.2 Impedance Characteristics

To utilize the Radio Manufacturers Association standard transmission lines having a characteristic impedance of 51.5 ohms, the input impendance at the junction of the four transmission lines in the center of the loop should be approximately 100 ohms because loops are always used in pairs. This 100 ohms is obtained by paralleling the four loop transmission lines, each of which has an input impedance of 400 ohms. The input impedance of 400 ohms as seen looking into the loop lines from the junction unit is obtained by two series transformer sections in each coaxial line. Experiments show that these sections are respectively of 100 ohms and 50 ohms impedance as seen from a loop corner to the junction at the center of the loop. The transformation that will insure the proper resistive component at the center of the loop can occur only at a single frequency; this frequency will be considered as being that at which the loop is resonant. It is important that the resonant condition of the loop be obtained approximately at a midband frequency. At operating frequencies on either side of the midband frequency, a tuning stub is inserted in parallel with the four transmission lines to adjust the impedance. To keep the stub to less than a quarter wavelength over the frequency range of 88 to 108 megacycles, it is necessary to use both open- and short-circuited stubs, as shown in Fig. 2.

1.3 Maximum Bandwidth Relative to Cost

It is well known that the input impedance characteristics of an antenna can be made less frequency selective by increasing the radius of the radiating elements. This could be accomplished by using large-diameter transmission lines in that portion forming the radiating sides of the loop. This becomes an expensive expedient, however, and it has been found that a square box enclosing a $1\frac{5}{8}$-inch outside diameter coaxial transmission line serves equally well. It must also be pointed out that when radiating elements having large radius are employed, the sides of the loop will undergo a further reduction in

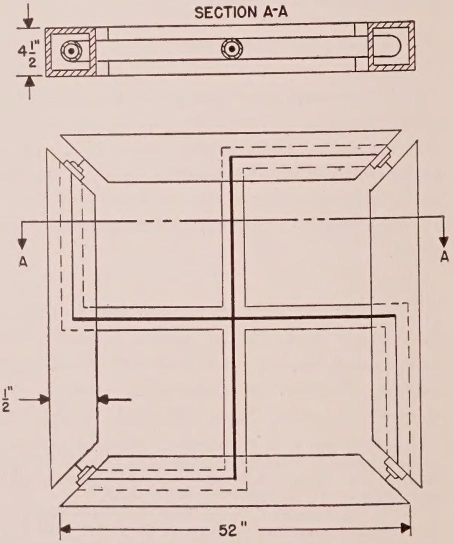

Fig. 3—Dimensions of the square loop.

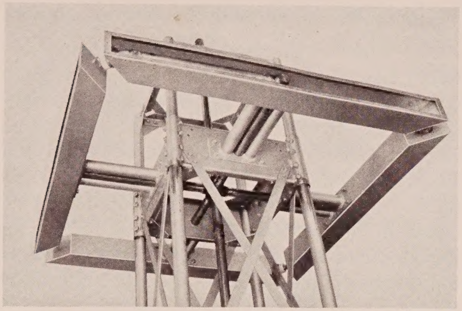

Fig. 4—Square-loop assembly with covers removed from radiating elements. Note connection of center conductor of coaxial line to adjacent section. The loop assembly is mounted on an integral section of lattice tower.

physical length because of the increased capacitance between the ends of adjacent elements. For the 88- to 108-megacycle range, the final dimensions of the loop incorporating the square-box assembly and the stub tuner are shown in Fig. 3.

Experiments show that the electrical characteristics are not materially altered by utilizing an area of two square feet within the loop to accommodate the supporting tower. The loop, mounted on a supporting structure, may be seen in Fig. 4. In the center of the loop may be seen the six-way junction unit, to which are connected the four coaxial lines feeding the loop, the tuning stub, and the transmitter line. The front cover plates of the radiating sections have been removed.

1.4 Minimum Number of Selective Circuits

By employing radiators of large diameter and of suitable length, the *Q* has been lowered to improve the input impedance characteristics between 88 and 108 megacycles. Radiators of these dimensions offer a *Q* low enough to make feasible the use of two impedance-transformation sections as mentioned above without seriously affecting bandwidth. By stacking the loops approximately a wavelength apart, the mutual coupling between elements has been greatly reduced. The radiation pattern for the ideal small loop shows minimum radiation in a direction normal to the plane of the loop. This characteristic, together with the wavelength spacing between loops, tends to reduce the number of coupled selective circuits.

Before going into the subject of stacking, consider a single square loop. The variation of input impedance with frequency at the center junction of a single loop is plotted in Fig. 5. In Fig. 6, the standing-wave ratio is given for a single loop with proper stub termination over the frequency range of interest. The standing-wave ratios are in the neighborhood of 2 to 1, thus making the resistive component about 100 ohms.

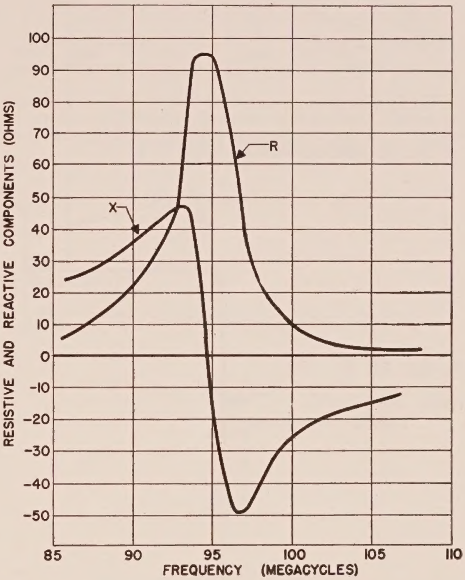

Fig. 5—Series input impedance—frequency characteristic of a single square-loop antenna.

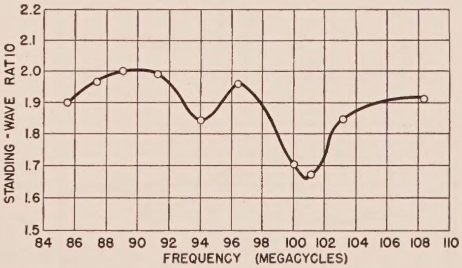

Fig. 6—Standing-wave ratio—frequency characteristic of a single square-loop antenna. Measurements are taken on a 50-ohm coaxial line.

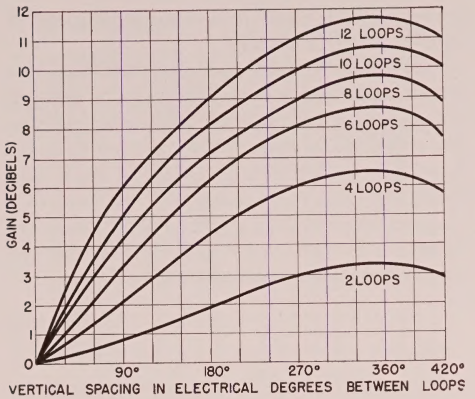

Fig. 7—The power gain of the square-loop antenna can be greatly increased by vertically stacking a number of loops. Note that the highest gain is obtained with one-wavelength spacings between loops.

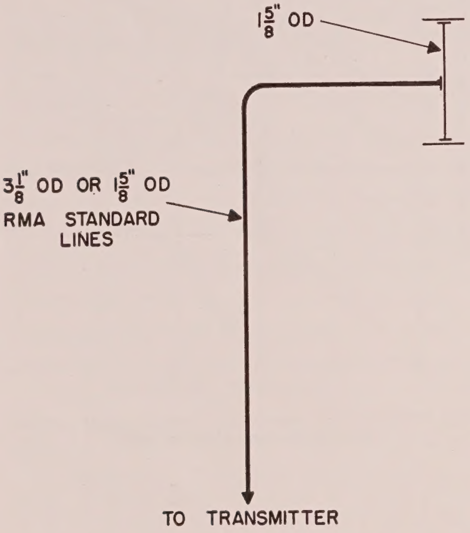

Fig. 8—Coaxial-line feed system for a two-loop array.

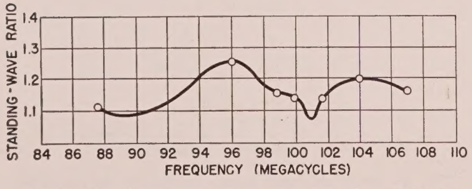

Fig. 9—Standing-wave ratio—frequency characteristic of the two-loop array (Fig. 8).

1.5 Stacking of Elements

Fig. 7 shows that the optimum spacing[3] between the loops is 360 degrees or one electrical wavelength. For this particular design, the wavelength chosen will be the geometric mean value in the frequency range of 88 to 108 megacycles. With two loops, one above the other, the transmission lines are connected as shown in Fig. 8, where the line connecting the two loops has an impedance of 100 ohms, thus requiring no transformation in that section of line. It is apparent that if connection is made at the mid-point of the 100-ohm section, the impedance seen will be 50 ohms, and the two loops will be in phase as a result of the equal lengths of transmission line. A measurement of standing-wave ratio on the 50-ohm line supplying power to the two-loop array is shown in Fig. 9.

Such a two-loop system connected to the source of power at the mid-point between the loops may be considered to be a basic array. Two such basic arrays may be used to provide additional gain, with the power being supplied through the transmission-line system shown in Fig. 10. The function of the transformation section is to match the 25-ohm impedance of the mid-point between the two basic arrays to the main 50-ohm line to the transmitter. This arrangement may be extended to provide for 8- and 16-loop arrays.

Arrangements for other than 2-, 4-, 8-, and 16-loop arrays will require modification of the standard transmission-line system. It must always be borne in mind that the transmission lines from any given feed point to the loops themselves must have equal lengths. A possible arrangement for a 6-loop array is shown in Fig. 11.

1.6 Transmission-Line Systems

The method of exciting the square loops has an important bearing on the horizontal radiation pattern, and consequently on the relative gain that can be obtained from an array. By feeding the four loop elements from a common point and making the electrical lengths of the transmission lines to their respective loop elements identical, proper phase and amplitude conditions are established to produce highest gain in the horizontal plane. Moreover, each loop element then receives its proportionate amount of power.

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[3] A. G. Kandoian, "Three New Antenna Types and Their Applications," *Electrical Communication*, v. 23, p. 32; March, 1940, and *Proceedings of the I.R.E. and Waves and Electrons*, v. 1, p. 74; February, 1946. Gain is related to radiation from a single small loop.

In designing a standard transmission-line system, maximum use is made of the Radio Manufacturers Association proposed standard 51.5-ohm transmission lines. Depending on the voltage breakdown requirements, these coaxial lines are available with outside diameters of either $1\frac{5}{8}$ or $3\frac{1}{8}$ inches.

It is beyond the scope of this paper to consider all the electrical and mechanical characteristics of a feed system. Many electrical factors, such as attenuation and phase velocity, can be calculated from a given insulator-bead separation within the coaxial line and a given consistency of bead characteristics. Likewise, in dealing with the mechanical problems, dead weight, wind loading with or without ice, as well as expansion and contraction, can be computed. When considering the problem of expansion and contraction it is necessary to know not only the ambient temperatures, but also the heating resulting from the radio-frequency currents within the transmission line itself. In discussing the over-all feed systems, it may be of interest to describe a few of the parameters that are not easily calculated, and yet are not difficult to measure. Consider, for instance, the insulator characteristics as an electrical problem, and expansion and contraction as a mechanical problem.

1.6.1 *Insulators*

When determining the optimum structure of a bead-type insulator for maximum flashover voltage, actual tests provide the best answer.

It is reassuring to make these tests at the frequency and power at which the insulator will operate, thus avoiding difficulties with factors which may be neglected in tests made under other conditions. All insulators under test were constructed of steatite L-5.

Table I gives an over-all picture of the various types of bead insulators used and their characteristics. These beads are illustrated in Fig. 12.

The low flashover voltage of types *A* and *B* spacer beads are no doubt caused by high voltage gradients at the point of tangency to the inner conductor. Concentrated heating evidently occurs at this point, followed by high ionization and flashover. The *E* type of bead insulator provides best voltage breakdown properties and will

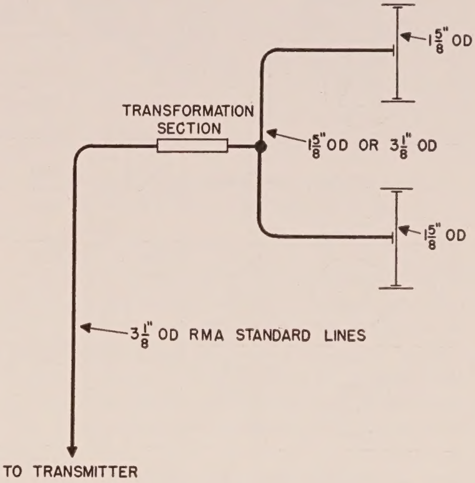

$1\frac{5}{8}$" OD
TRANSFORMATION SECTION
$1\frac{5}{8}$" OD OR $3\frac{1}{8}$" OD
$1\frac{5}{8}$" OD
$3\frac{1}{8}$" OD RMA STANDARD LINES
TO TRANSMITTER

Fig. 10—Feed arrangement for a four-loop array.

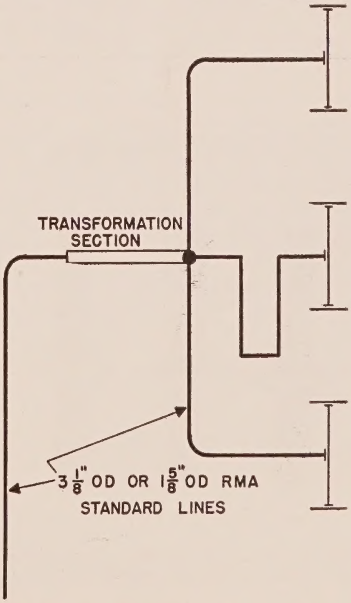
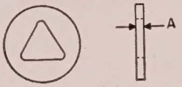
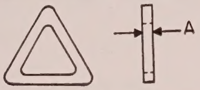
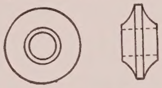
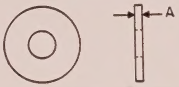
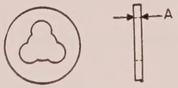


Fig. 11—Feed system for a six-loop array.

TABLE I

Characteristics of Coaxial Insulator Beads ($f = 108$ Mc)

Sample		Thickness A Inches	Voltage (RMS)			Observed Effect	Stabilized Temperature Degrees Centigrade	
			Minimum	Up to 5 Seconds Momentary	Duration		1400 V.	2100 V.
A		1/4	Minimum expected service voltage is 1400 at 10 kilowatts with $1\frac{5}{8}$-inch transmission lines.	2450 to 2750		Breakdown tangent to inner conductor	43.5	No data
		3/16		1900 to 2300			41.0	No data
B		1/4		3200		Breakdown tangent to inner conductor	39	No data
		3/16		3200	15 sec. at 3400		38	No data
C		Variable		4600	90 sec. at 4000 180 sec. at 3500	Breakdown tangent to inner conductor	No data	No data
D		3/16		2890 to 3270	Instantaneously flashed over at 3270	Breakdown occurs radially across inner and outer diameters	No data	No data
E		1/4		Momentary flashover above 5000	10 sec. at 4000	Breakdown begins at a corner of cutout and then fills the cutout.	No data	No data
		3/16		Momentary flashover above 5000	90 sec. at 4000 180 sec. at 4500		32	50

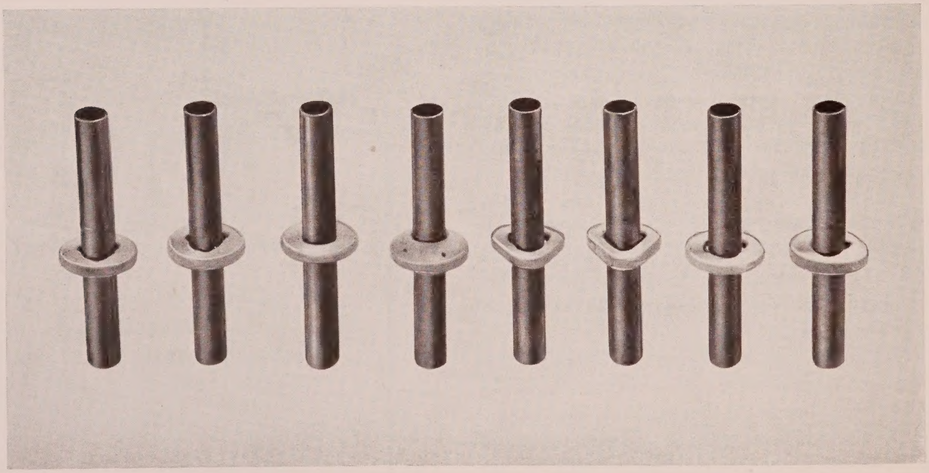


Fig. 12—Eight types of bead insulators for coaxial lines carrying high powers in the frequency-modulation band.

be used wherever the transmission-line characteristics permit. It might be noted that the use of this cloverleaf design is not contemplated in the 100-ohm transmission lines that connect two or more of the basic two-loop arrays. The simple disk type will be used in these sections because of the difficulties involved in cutting the cloverleaf inner hole for a $\frac{1}{4}$-inch conductor.

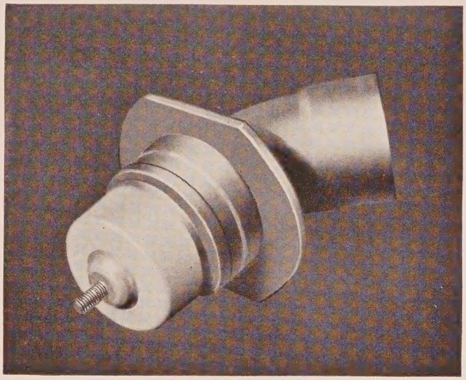


Fig. 13—End-seal insulator used at each corner of the square loop. This insulator must have proper electrical characteristics and must also prevent gas leakage.

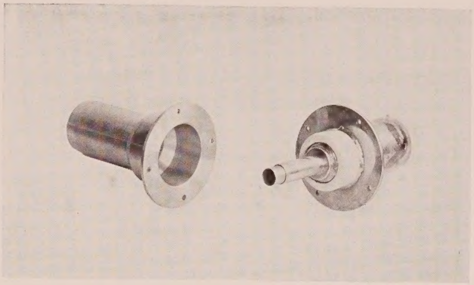


Fig. 14—Gas-sealing insulator for transmitter end of $1\frac{5}{8}$-inch, 50-ohm coaxial transmission line.

Table II gives test results on two end-seal insulators and on an anchor bead insulator. Fig. 13 shows the end-seal unit *F* used at the loop corners. The other end-seal insulator is used at the transmitter termination of the line and may be seen in Fig. 14. These two units must be gas tight and of such construction as to avoid line discontinuities. They must withstand high voltages without flashing over.

The anchor insulator *H* will support the inner conductor for a vertical run of at least 100 feet. Holes are provided to allow free flow of gas across the bead. A cross-sectional drawing is given in Fig. 15.

TABLE II

CHARACTERISTICS OF COAXIAL LINE INSULATOR FITTINGS

Sample	Minimum Expected Service Voltage	Momentary Voltage	Observed Effect	Pressure-Under-Temperature Cycle		
				Time	Temperature Degrees Centigrade	Pressure Pounds Per Square Inch
F End Seal "A" 100 Ω	2500	3200	Breakdown at rear head only. Not at end seal proper.	0	3	75
				23 Mins.	−35	64
				Removed unit from refrigerator, pressure returned to 75 PSI after 1/2 hour.		
				0	26	38
				5 Hours	105	50
				Removed unit from oven, pressure returned to 37.5 PSI at room temperature. End seal removed at 0 PSI. Ring was inspected and found to be satisfactory.		
G End Seal "B" 50 Ω	1400	4900	End seal did not break down at a voltage of 4900 in 5 minutes.	No data		
H Anchor Bead	1400	2720 to 3110	Anchor bead breakdown occurred radially across inner and outer diameters	Not required.		

1.6.2 Expansion Joints

Temperature changes are related to the weather and to the heating effects of the radio-frequency currents in the transmission line. They may be compensated for by the use of expansion joints. Two types of expansion joints have been designed for the outer conductor; the single-bellows and the double-bellows types. The single-bellows type, Fig. 16, is designed for the $1\frac{5}{8}$- and $3\frac{1}{8}$-inch lines, and will be used in the sections feeding the square loops. The double-bellows type is designed for the main vertical runs and is capable of taking up the expansion and contraction of 100-foot lengths of transmission line. As both the anchor insulator and expansion joint are required on long vertical runs, these two units are incorporated in one assembly. Calibrated expansion markings are provided so that the proper relaxation setting may be made for the temperature at the time of installation. After installation has been completed, one end of the guard cylinder over the bellows is released, thus allowing free movement of the joint. The inner-conductor expansion joints for all sizes of lines are of the "banana" type.

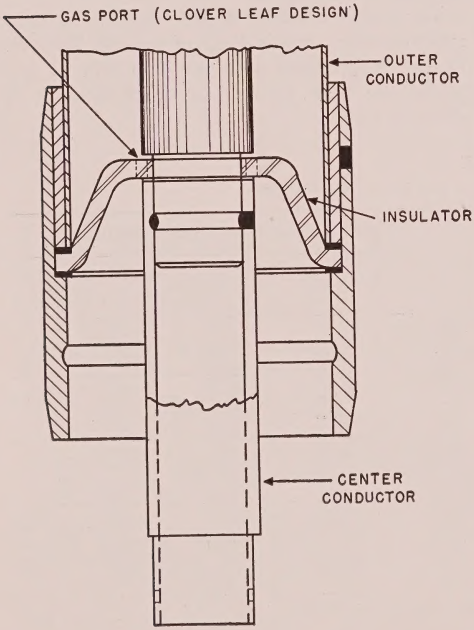


Fig. 15—Cross-sectional view of an anchor insulator which will support the inner conductor of a vertical 100-foot length of coaxial line. This unit is incorporated in a single assembly with a double-bellows expansion joint.

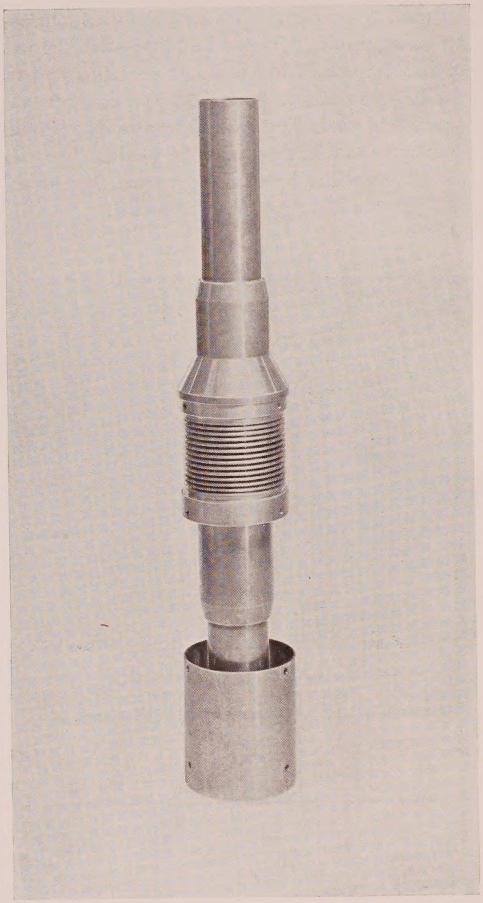


Fig. 16—Single-bellows expansion joint. This unit is used in short sections of coaxial line to absorb expansion and contraction resulting from temperature changes.

In the over-all expansion problem, it is also necessary to consider the characteristics of the supporting tower. The curves in Fig. 17 depict the expansion and contraction on the 100-foot vertical spans. These curves are based on average weather conditions and temperature rise resulting from radio-frequency heating. In making these assumptions, it is necessary that the radio-frequency heating be considered uniform throughout the entire length for both the inner and outer conductors. The curves in Fig. 18 show the heat-expansion characteristics of the $1\frac{5}{8}$-inch Radio Manufacturers Association line with bead separations of 12 inches. These measurements were conducted at a frequency of 108 megacycles with standing-wave ratios from 1:1 to 4:1, and a power of about 10 kilowatts. It is of interest to see the effect of the bullet connector which is a solid brass stud soldered to the inner conductor to make contact between the ends of adjacent inner conductors.

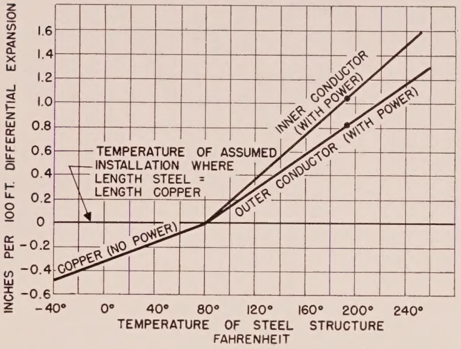


Fig. 17—Expansion of a 100-foot vertical span of coaxial line mounted in a steel tower. The copper and steel have different temperature coefficients of expansion, resulting in a change in the length of the line with respect to the length of the tower with temperature variations. Heating resulting from power transmitted through the line must also be considered. The installation of expansion joints prevents the line from buckling.

Fig. 19 shows a typical 8-loop array. The expansion joints marked *C*1 are of the single-bellows type, and those marked *C*2 are the composite double-bellows type with anchor insulator.

2 Mechanical Problems

2.1 Ease of Installation

From the curves in Fig. 7, it is evident that the antenna power gain is not seriously affected as the vertical loop spacing is varied around the 360-degree point. Advantage is taken of this characteristic by constructing the square loops on a lattice tower structure of fixed length. Thus, by choosing a spacing of 9 feet 3 inches, the

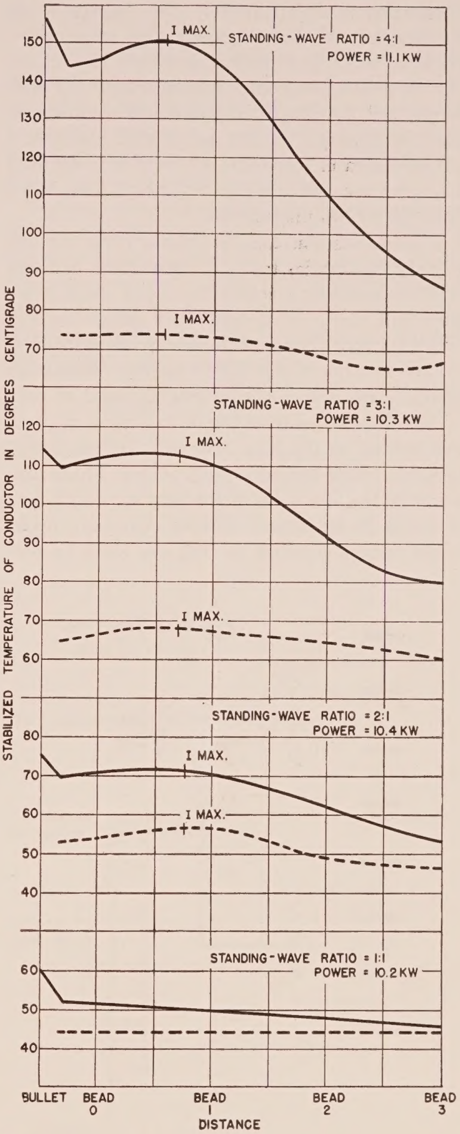


Fig. 18—Effect of standing waves on the temperature of the inner and outer conductors (shown solid and dashed, respectively) at various points along a coaxial line. Temperatures are highest at the current loops because of the greater *IR* losses. These measurements were taken on a $1\frac{5}{8}$-inch, 50-ohm line; the spacing between the insulator beads is 12 inches; the ambient temperature 25–30 degrees centigrade, and the frequency 108 megacycles.

separation is approximately 300 degrees at 88 megacycles, and 366 degrees at 108 megacycles. Such a two-loop structure is shown in Fig. 20. As very little mutual coupling exists between loops, and the proper impedance characteristics may be obtained by the use of stub sections, a single design of loops and spacings may be used to cover the entire frequency-modulation band from 88 to 108 megacycles.

Further standardization results from the use of only two lattice tower designs. Each of these will support four stacked loops, but the heavier unit will carry in addition the lighter section (with its four loops) to make up an eight-loop array. The two types have identical fabrication characteristics and can be interchanged at will.

As will be noted from Fig. 4, two rigid supports are welded to the loop assembly and anchored securely to the lattice superstructure. These supports on the four sides of the loop have sufficient strength to withstand stresses that may occur while personnel climb around the loops in performing such maintenance jobs as replacing obstruction lamps or bleeding gas from the transmission lines.

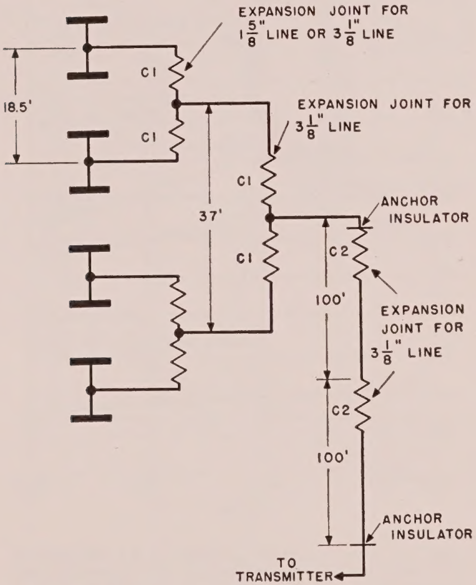

EXPANSION JOINT FOR 1 5/8" LINE OR 3 1/8" LINE
C1
C1
18.5'
EXPANSION JOINT FOR 3 1/8" LINE
C1
37'
C1
ANCHOR INSULATOR
C2
100'
EXPANSION JOINT FOR 3 1/8" LINE
C2
100'
ANCHOR INSULATOR
TO TRANSMITTER

Fig. 19—Coaxial-line arrangements for feeding an eight-loop array. Expansion joints marked *C1* are the single-bellows type; those marked *C2* are of the double-bellows structure. All horizontal sections are extremely short.

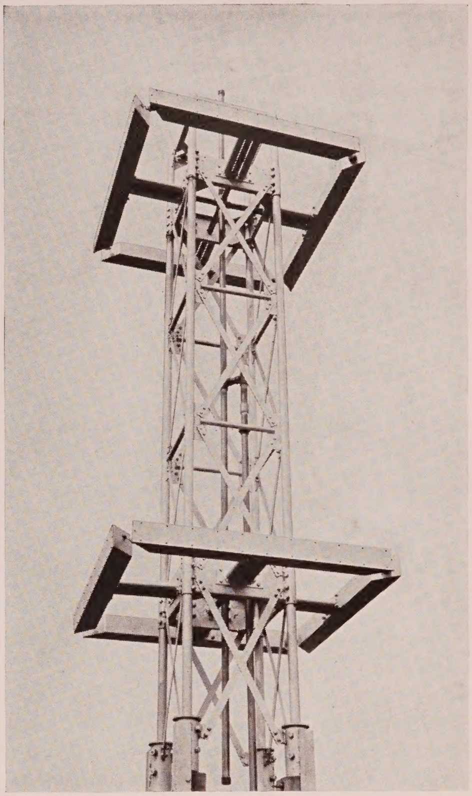


Fig. 20—Array of two square loops showing steel lattice tower sections. Note the coaxial feed lines in the center of the tower and the reactance-matching stub attached to the line junction in the center of each loop.

The coaxial feed system may be prefabricated, and thus presents a relatively easy installation operation. The use of solderless fittings likewise adds to the ease of over-all assembly. The tower structure has ample room in the interior for two $3\frac{1}{8}$-inch transmission lines, or one $6\frac{1}{8}$-inch line. With these line sizes, it is possible to feed well over 50 kilowatts into the array.

The entire antenna structure is made in accordance with the latest Radio Manufacturers Association standard recommendations.

2.2 MAINTENANCE

Every effort has been made in mechanical as well as electrical design to minimize maintenance requirements. Although heating elements are not supplied with the standard loops, it is a relatively easy matter to insert heating elements under the covers when weather conditions demand ice and sleet melting.

The complete transmission-line assembly is gas filled, and all fittings are designed to insure gas tightness. Valves inserted at the corners of each loop allow the bleeding of gas.

2.3 COST

The use of a single design to operate anywhere in the transmission band allocated to frequency-modulated broadcasting will reduce costs to a reasonable minimum. A relatively large stock of standard elements may be maintained economically and be available at all times.

A prospective user of eight loops who contemplates immediate use of four loops may obtain the four heavier lattice sections, and procure a supporting tower capable of taking the entire ultimate load. The later addition of four lighter sections placed atop the heavier units readily provides an eight-element array.

Triode Amplification Factors

By J. H. FREMLIN, M.A., Ph.D., F.Inst.P., A.M.I.E.E.,
Nuffield Research Fellow, University of Birmingham, Birmingham, England

R. N. HALL, AND P. A. SHATFORD
Standard Telephones and Cables Ltd., London, England

DURING the past few years the importance in valve design of obtaining a high ratio of mutual conductance to anode current at a practical value of slope has been steadily increasing. Such high ratios can be obtained in several ways:

Increase of working electrode area for a fixed total current will give the desired result, in the same way as would putting a number of valves in parallel and biasing them back.

Secondary-emission multiplication can be used to get a practical value of slope at ratios approaching the theoretical limit fixed by cathode temperature; at present in the region of 11 milliamperes per volt per milliampere.

Finally, the relative electrode positions can be changed, and in particular the cathode-to-control-grid clearance can be decreased.

The first method is simple but involves an even more rapid increase in interelectrode capacitances, which is undesirable for a large range of applications.

The second involves considerable complexity of electrode design, and moreover leads to long electron transit-times which again interfere with operation at very high frequencies.

The third has neither of these disadvantages, although if pressed very far it leads to mechanical difficulties. This is due in large part to the fact that it is essential to keep the electric field at the cathode surface due to the charged grid approximately uniform. The distance between cathode surface and grid plane must then not become much less than the pitch of the grid wires (grid pitch). Hence, a very large number of turns per unit length of grid will be needed. The difficulty of handling the grid wire increases rapidly as the wire becomes thinner, hence it is usual in many high-slope valves nowadays to use larger values of "shadow ratio" (grid-wire diameter to grid pitch) than would ever have been considered a few years ago. It is the main object of this paper, after a brief discussion of the theoretical formulae available, to describe some experimental results which were obtained to assist design work with grids of high shadow ratio. Tables are given from the formula, shown experimentally to be the best, for convenient use in calculating amplification factors for valves using such grids. Theoretical and experimental results showing the effect of abnormally close electrode spacings will also be described.

Valves using parallel plane electrodes will alone be considered.

• • •

1 *Thick Wires and Formulae for Amplification Factors*

In this section we shall consider the validity of various formulae as a function of the ratio of wire diameter to grid pitch. The further complications introduced if the grid approaches other electrodes will be treated in the next section.

It is well known that the early formulae due to Schottky and Miller[1] and to King, are only roughly valid for thin grid wires. If we call grid-wire diameter d, number of turns per unit length of the grid N, and distance between anode surface and the plane through the centres of the control-grid wire S, we may write Miller's formula for the amplification factor of a parallel-plane triode as

$$\mu = \frac{2\pi NS}{-\log_e\left(2\sin\frac{\pi Nd}{2}\right)}. \qquad (1)$$

This gives an infinite value for μ when $\pi Nd/2 = \pi/6$, or when the shadow ratio Nd is 0.33, and must clearly give much too large results over a considerable range of shadow ratios smaller than this. Even at $Nd = 0.20$, the formula gives a result too large by about 16 per cent.

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[1] J. M. Miller, "The Dependence of the Amplification Constant and Internal Plate Circuit Resistance of a Three-Electrode Vacuum Tube Upon the Structural Dimensions," *Proceedings of the I.R.E.*, v. 8, pp. 64–74; February, 1920.

The improved formula given by Vogdes and Elder,[2] which for parallel plane electrodes is

$$\mu=\frac{2\pi NS}{\log_e \coth \pi Nd}-\frac{\log_e \cosh \pi Nd}{\log_e \coth \pi Nd}, \tag{2}$$

errs in the opposite direction, so that μ appears to have a finite value even when $Nd=1$ (the grid is a continuous metal surface). The error is appreciably less than that of (1) when the shadow ratio Nd is 0.2, for example, the value obtained for μ will be about 6 per cent too low, as against Miller's 16 per cent too great, and will still be within about 20 per cent of the correct value at $Nd=0.33$, where Miller's formula prophesies an infinite result.

More recently, formulae have been put forward by Ollendorff[3] and later still by Herne[4] which give results for which very much better accuracy is claimed.

Ollendorff's formula may be written

$$\mu=\frac{NS}{T}-\frac{\Delta}{T}, \tag{3}$$

where N and S have the same significance as in the earlier formulae, and T and Δ are quantities depending only on the shadow ratio Nd. Ollendorff claims that his method of calculation is capable of giving accurate results up to $Nd=1$. His basic equations are not suitable for numerical substitution, however, and he gives a table of four successive approximations from which T and Δ can be calculated. The first approximation is identical with King's formula

$$\mu=\frac{2\pi NS}{-\log_e \pi Nd},$$

which has nearly the same range of validity as Miller's (1), while the fourth approximation is claimed to be workable up to a shadow ratio of 0.9. This is not the case; the value of μ calculated from it goes to infinity at $Nd=0.797$, but the rapid convergence of the successive approximations for values of Nd approaching 0.4 make it seem likely to be much superior to previous formulae.

The fourth approximations given by Ollendorff for Δ and T, respectively, are as follows, putting

$$y=\frac{1}{3}\left(\frac{\pi Nd}{2}\right)^2, \tag{4}$$

$$\pi\Delta_4=\frac{3y}{1+y-\dfrac{0.21y^4}{\dfrac{7}{4}+y^3}}, \tag{5}$$

and

$$2\pi T_4=-\log \pi Nd+\pi\Delta_4-\frac{y^2}{2+1.2y^2}. \tag{6}$$

(It may be noted that the numbers 0.21 and 1.2 are exact, not approximate.)

Herne's formula in our nomenclature is

$$\mu=b(2\pi NS-c), \tag{7}$$

where b and c are quantities that depend on the shadow ratio only, for which Herne gives a table. It will be seen that this equation is in form similar to those of Vogdes and Elder, and of Ollendorff; Herne's quantity b corresponding to Ollendorff's quantity $1/2\pi T$, and Herne's c to Ollendorff's $2\pi\Delta$.

The numerical difference between the results obtained from the two formulae is not significant for most of the range which has so far been of practical use, but from the table given it appears that Herne's formula causes the calculated value of μ to go to infinity at a slightly smaller shadow ratio than does Ollendorff's.

2 Limits of Validity as the Grid Approaches Cathode or Anode

The ranges of validity of all of the above formulae are limited by the assumption that the cathode-grid and anode-grid distances, l_g and S, respectively, are large compared to the grid pitch a (or $1/N$).

It has been shown[5] that for moderately thin wires, the effect of reduction of l_g/a, or Nl_g, becomes of practical significance when l_g is about $0.8a$. The formula derived, making the same approximations as to wire thickness as did Miller, is

$$\mu_0=4\pi N\frac{l_g-l_a \tanh \pi Nl_g}{4\pi Nl_g \tanh \pi Nl_g-\log_e\left[1+\dfrac{\sinh^2 2\pi Nl_a}{\sin^2 \dfrac{\pi Nd}{2}}\right]}, \tag{8}$$

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[2] F. B. Vodges and F. R. Elder, Physical Review, v. 24, p. 683, 1924.

[3] F. Ollendorff, *Electrotechnik und Maschinenbau*, v. 52, p. 585, 1934.

[4] H. Herne, *Wireless Engineer*, v. 21, p. 59, February, 1944.

[5] J. H. Fremlin, "Calculation of Triode Constants," *Electrical Communication*, v. 18, pp. 33–49; July, 1939.

where l_a is the distance between anode and cathode ($l_a = l_g + S$), and where μ_0 is the value of μ near out-off; in this case μ increases with current. Curves showing the increase are given in a reference.[5]

For values of l_g not less than about $0.4a$, this may be simplified to

$$\mu_0 - \mu_M \left\{ 1 + 2e^{-2\pi N l_g} \left[1 + \frac{l_g(\mu_M + 1)}{S} \right] \right\}^{-1}, \quad (9)$$

where μ_M is the value of μ given by Miller's formula (1).

It is probable that the "variable μ" effect will be somewhat greater than is suggested by (9) for thicker wires, but this will be a second-order effect, though it may be desirable to replace μ_M by the value of μ found from Ollendorff's formula (3).

The effect of reducing the anode-grid distance while keeping the cathode-grid distance large has been discussed by Salzberg.[6] He has developed a formula for μ under these conditions, which for the case of the plane triode is

$$\mu = \frac{2\pi N_s - \log_e \cosh \pi N d}{-\log_e \tanh \pi N d + \log_e (1 - e^{-4\pi N a} \cosh^2 \pi N d)}. \quad (10)$$

This was developed using the same approximations as did Vogdes and Elder,[2] and is thus, presumably, valid up to a similar value of shadow ratio.

Before Salzberg's paper was available, another formula dealing with the same conditions was developed by one of the authors using Maxwell's expressions for the potential distribution due to a charged grid plus that due to its image in the anode plane, the method being identical with that used in a reference[5] to treat the case of small cathode-grid distance. The formula obtained may be written

$$\frac{1}{\mu} = \frac{1}{\mu_M} + \frac{1}{4\pi N S} \log_e \times (1 - 2e^{-4\pi N S} \cos \pi N d + e^{-8\pi N S}), \quad (11)$$

where μ_M is the value of μ obtained by Miller's formula (1).

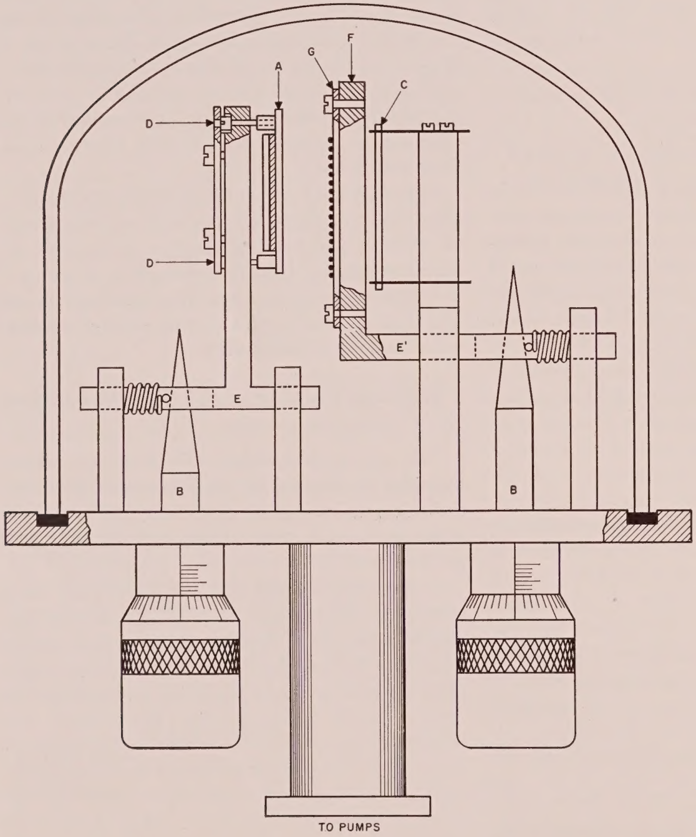

Fig. 1—Movable-electrode triode. *A*—**anode structure;** *B, B*—**tapered micrometer drives;** *C*—**fixed cathode;** *D*—**anode fine-adjustment screws;** *E, E'*—**sliding carriages;** *F*—**grid support;** *G*—**plate on which grid is mounted.**

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[6] B. Salzberg, "Formulas for the Amplification Factors of Triodes," *Proceedings of the I.R.E.*, v. 30, pp. 134–137; March, 1942.

This is likely to become inaccurate before (10) as the grid-wire diameter is increased but is given here because it was used for comparison with experimental measurements which may be worth quoting, as they were obtained by a method different from that used by Salzberg.

The form of (11) is also convenient in that we can replace $1/\mu_M$ by $1/\mu_{OU}$, when μ_{OU} is the value given by Ollendorf (3) which, though perhaps difficult to justify logically, gives results which are believed likely to be better than those of either (10) or (11). The difference between (10) and (11) is not appreciable either for the experimental system used by Salzberg or for that described below.

It may seem curious at first sight that the value of μ obtained is greater than those worked out from formulae which neglect the nonuniformity of charge on the anode. The reason for this may perhaps be visualized by considering that the uniform charge on the anode assumed by Miller will tend, as the grid approaches the anode, to collect mainly behind the grid wires where its effect on the cathode field is least. The formula is more sensitive even than Miller's to the effect of increasing wire thickness, although this effect is not very great. When the anode-grid distance is 0.2 of the grid pitch for example, the calculated value of μ goes to infinity at $Nd=0.326$ instead of at $Nd=0.333$.

3 Experimental Determination of Amplification Factor for Triodes with Thick Grid Wires

A continuously pumped system was constructed in which the anode-grid and grid-cathode distances could be varied by micrometer heads (Fig. 1). These heads drive tapered rods *B*, *B* against which carriages *E*, *E′* are pressed by springs, thus giving carriage movements at right angles with a 5:1 reduction. The anode can be reset at any position to within 0.02 millimeter, and the grid-cathode spacing to within 0.05 millimeter.

The anode *A* was mounted on carriage *E* by three 10 British Association screws *D*, giving a fine adjustment for setting the anode parallel to the grid.

The anode structure consisted of a guard ring of molybdenum sheet 0.4 millimeter thick having a slot 12.7 millimeters long and 1.52 millimeters wide for the anode.

The anode proper was another sheet of molybdenum with a projecting ridge 11.7 millimeters long and 1.42 millimeters wide, which was riveted to the guard ring through mica insulators (Fig. 2).

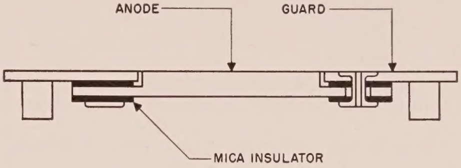

Fig. 2—Anode structure.

After riveting, the faces of both anode and guard ring were ground flat so that there were no irregularities in the surface, except the 0.05-millimeter slot between anode and guard ring.

The grids were constructed (except the one of largest shadow ratio) by making frames *G* (Fig. 1) of 0.635-millimeter steel plate which could be bolted to the grid support *F*. The wires were welded to the face of these plates with the help of an auxiliary frame consisting of threaded rods of 1-millimeter pitch on which the wire was first wound and then stretched.

When measured under a travelling microscope, the deviation from the nominal pitch was found to be never greater than 0.02 millimeter (Table I).

TABLE I
MEASUREMENTS OF UNIFORMITY OF GRID PITCH

Travelling Microscope Readings in Millimeters	Distance Between Successive Wires in Millimeters	Running Average of 7 Pitches
63.48	0.98	—
62.50	1.00	—
61.50	1.00	—
60.50	1.02	—
59.48	1.00	—
58.48	0.99	—
57.49	1.01	—
56.48	1.00	7.00
55.48	0.98	7.02
54.50	1.00	7.00
53.50	1.02	7.00
52.48	0.98	7.00
51.50	1.00	6.98
50.50	0.90	6.99
		Total: 48.99

The diameters of the grid wires were checked by a comparator which measured the difference between each wire diameter and the thickness of a standard feeler gauge. A sample set of readings is shown in Table II.

TABLE II

Deviation of Wire Diameter from Standard (0.5076 mm.)

0.000	−0.002	−0.002
−0.001	−0.002	+0.003
−0.001	−0.002	+0.002
−0.003	−0.003	0.000
−0.003	−0.002	−0.002
−0.003	−0.002	−0.002

Average deviation = −0.0014 millimeter.
Average diameter = 0.5076 − 0.0014 = 0.5062 millimeter.

For the grid of largest shadow ratio, a different method had to be used. A pitch of 1 millimeter would have necessitated a wire diameter of about 0.71 millimeter, and the only wire of that diameter obtainable was not round; its diameter varied by +0.000 to −0.012 millimeter. The pitch was therefore reduced to that of a 3 British Association screw (0.73 millimeter), and wire of 0.508-millimeter diameter was used.

Welding the wire to a steel frame was not feasible, however, and the grid was wound directly on a frame in which the two uprights consisted of steel 3 British Association screws.

The grid was mounted on the carriage E' (Fig. 1), and the cathode-grid distance adjusted by moving the grid until it just touched the cathode C, as indicated by an electrical contact between the two. It was then moved the required distance as determined by the micrometer head. In each case the grid-cathode distance was 1.2 times the grid pitch.

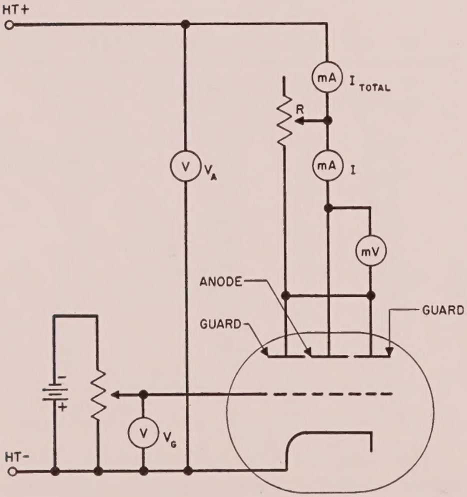

Fig. 3—Circuit Diagram.

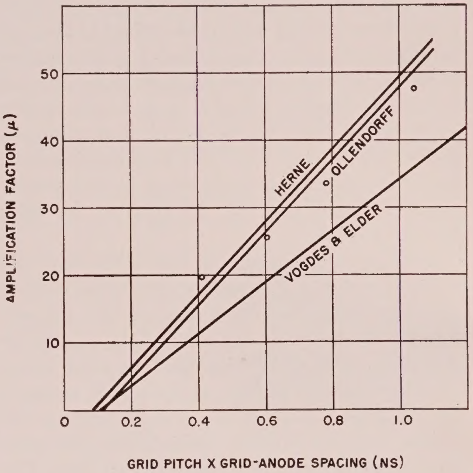

Fig. 4—Experimental and theoretical variation of μ with grid-anode spacing for a shadow ratio $(Nd) = 0.394$. The lines are calculated from the formulae of the authors named, while the points represent experimental values.

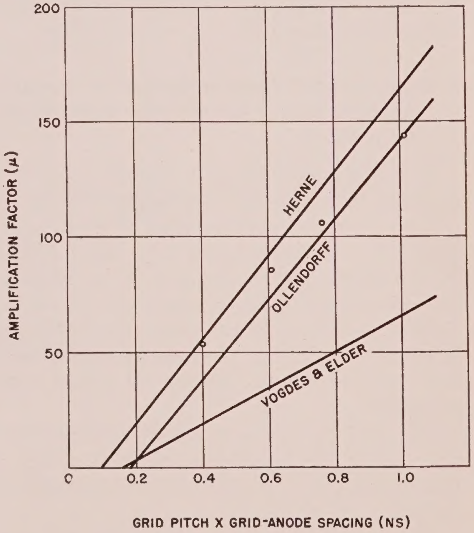

Fig. 5—Experimental and theoretical variation of μ with grid-anode spacing for a shadow ratio $(Nd) = 0.505$.

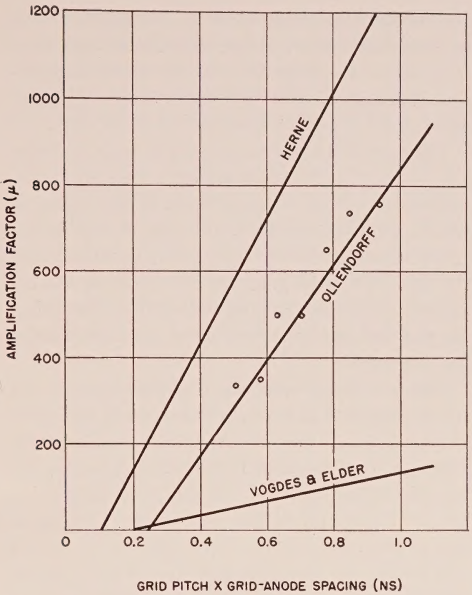

Fig. 6—Experimental and theoretical variation of μ with grid-anode spacing for a shadow ratio $(Nd)=0.635$.

The cathode consisted of a nickel tube flattened to give a surface approximately 6.3 millimeters wide, and the coated surface was 15.2 millimeters long.

The circuit used is shown in Fig. 3. The variable resistance R was adjusted for each setting of V_g to balance out any difference of potential between the anode and guard ring. Balance was indicated by zero deflection of the millivoltmeter (mV).

The anode-to-grid distance was checked after each series of experiments by setting the micrometer head to the readings which had been used, and inserting feeler gauges. The accuracy to be expected from this method is ±0.013 millimeter, assuming that no distortion due to thermal expansion takes place.

The results of these experiments are given in Figs. 4, 5, 6, and 7 for four different values of shadow ratio from about 0.4 to 0.66. In each chart the theoretical variation of μ with the anode-grid distance S is plotted for the formulae of Ollendorff, Herne, and Vogdes and Elder; the experimental values are given. Each value of μ

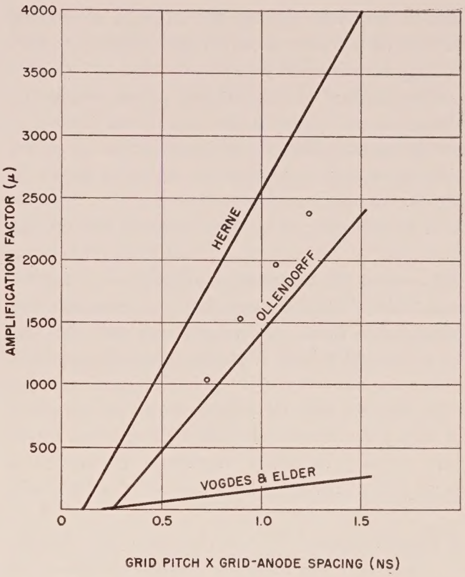

Fig. 7—Experimental and theoretical variation of μ with grid-anode spacing for a shadow ratio $(Nd)=0.665$.

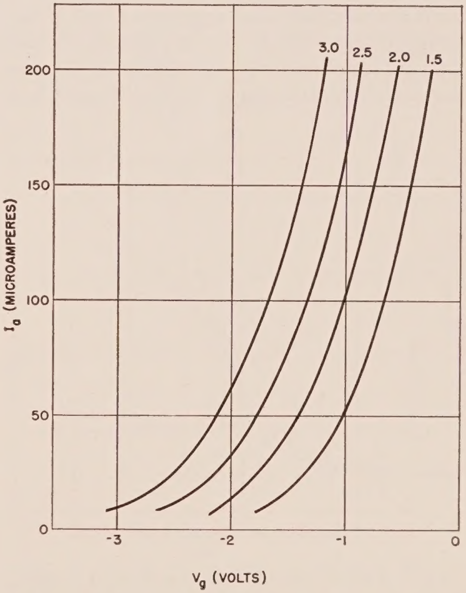

Fig. 8—Anode current-grid voltage for different values of anode voltage V_a in kilovolts. The following dimensions were employed: Shadow ratio $(Nd)=0.665$; grid pitch $(N)=0.73$ millimeter, cathode-grid spacing $(l_g)=0.85$ and, anode-grid spacing $(S)=0.67$.

was obtained by taking the average of several curves of anode current plotted against grid voltage for different anode potentials.

This method of determining μ was employed, rather than a bridge method, as it was desired to obtain characteristics on current and slope as well as on μ. A typical set of curves is shown in Fig. 8.

It can be seen (Figs. 4, 5, 6, and 7) that the formulae of Ollendorff and Herne agree with experiment far better than does that of Vogdes and Elder. The difference between the first two formulae is small, but experiment may be said to favor Ollendorff slightly, especially as both formulae will eventually give too large a value of μ; the fact that the experimental results given in Fig. 7 are above Ollendorff's line must therefore almost certainly represent experimental error due to the very large values of μ involved.

4 *Experimental Determination of Amplification Factors for Small Anode-Grid Spacing*

The practical requirements of this measurement are very different from those of the measurements described in the last section. In the previous experiments it was desired to measure amplification factors up to over a thousand, but the mechanical distances to be determined were of reasonable size; anode movement extended over about 1 millimeter, and the measurable resetting error of 0.02 millimeter would give an average error in μ of the order of 4 per cent.

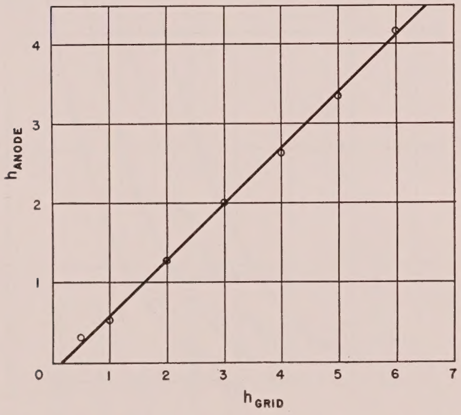

Fig. 9—Results obtained in the measurement of amplification factor on the rubber-sheet model. Grid shadow ratio (Nd) = 0.05, grid-cathode distance (l_g) = 1.6 times grid pitch (N), anode-grid distance (S) = 0.2 times grid pitch (N).

Here amplification factors will be near to unity, but the anode-grid distance for a grid pitch of 1 millimeter will be so small that, in the region in which we are interested, an error of 0.02 millimeter would give an error in μ of 10 to 40 per cent. As this error is as great as the whole of the expected deviation among well-known formulae, the method used previously was considered to be impracticable.

The method employed, therefore, was to set up a model of a triode section on a stretched rubber sheet. The way in which amplification factor can be measured with such a model has already been described.[5] Briefly, the inclination to the horizontal of the rubber sheet, when it leaves the edge of the strip which in the model represents the cathode, will correspond to the electric field intensity at the cathode surface. Vertical movement of grid or anode models corresponds to changes of potential on the grid and anode. The inclination of the sheet can accurately be followed by means of a small mirror attached to the sheet which reflects a beam of light to a distant surface. The ratio of movement of the anode model to that of the grid model required to keep the light beam stationary is then the amplification factor.

This system would have been entirely unsuitable for getting the results described in the last section, owing to the very large amplification factors involved. For mechanical simplicity the rubber-sheet model obviously works best for μ values not too far from unity. On the other hand, if the pitch of the grid model is, for example, 5 inches, it is easy to measure the grid-anode distance with adequate accuracy down to a very small fraction of the pitch.

For the experimental work, a value of shadow ratio Nd of 0.05 was chosen; first, because this would give $\mu=1$ at a suitable point of the range of anode distances used (when NS was about 0.33), and second, because the error to be expected (due to the assumption that the wires were thin) was small. The difference between the results given by Miller's and by Ollendorff's formulae was only about 0.2 per cent at $NS=1.0$, rising to 2 per cent at $NS=0.1$.

A typical set of measurements of corresponding grid and anode positions (h_a and h_g) for a fixed cathode field is shown in Fig. 9. The heights relative to the sheet are measured in terms of turns of the threaded rods used for the adjustments.

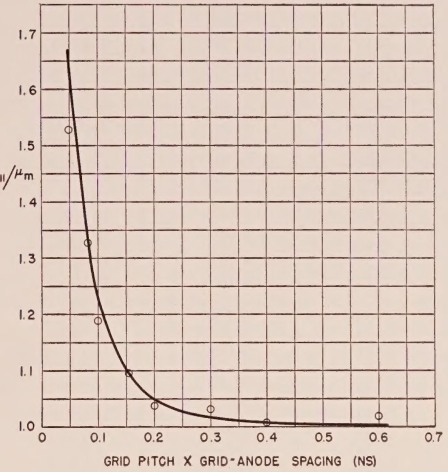

Fig. 10—Variation of amplification factor with grid-anode spacing. The full line represents μ_{11} in terms of μ_M, where μ_{11} is the value given by equation (11), and μ_M is Miller's value. The points marked indicate values found experimentally with the rubber sheet.

In Fig. 10, the ratio of μ to μ_M is plotted as a function of NS, the relative grid-anode spacing. The curve represents (11), and the circles represent the points determined by experiment. It will be seen that the experimental results follow the curve closely to $NS=0.075$, but that an appreciable deviation occurs at the smallest experimental point of NS. This is hardly surprising as at this point the distance of the surface of a wire from the anode is only the radius of the wire, and the assumption that the wires are thin is not to be expected to be fully justified. It seems clear that this deviation is unlikely to be of practical importance.

5 *Conclusion*

As a result of the experimental work here described, it is believed that for all thicknesses of grid wires likely to be met in practice, Ollendorff's formula may be used with confidence as long as the grid is sufficiently far from either cathode or anode. The formula is too complex to be conveniently used directly in valve design. Table III of the functions T and Δ involved in equations (5) and (6) has, however, been constructed in terms of the shadow-ratio Nd, which reduces the calculation of μ to a matter of two slide-rule operations.

The expression given for the case where the grid closely approaches the anode (11) has been shown by the use of the rubber sheet to be satisfactory if the grid wires are thin. Equation (11) can be written

$$\frac{1}{\mu}=\frac{1}{\mu_M}-y,$$

where μ_M is the value of μ obtained from Miller's formula (1). The correction term y, which again is too complicated for easy calculation, is graphed

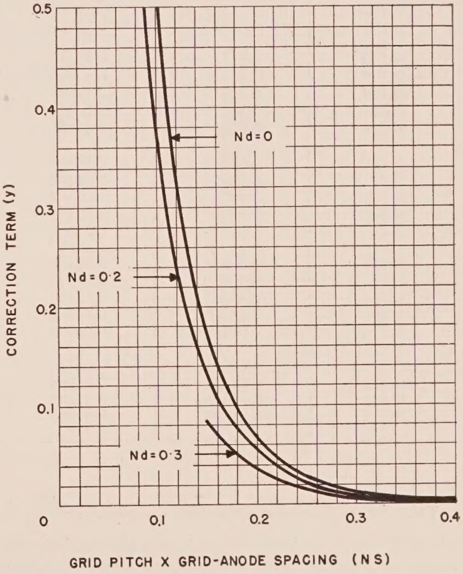

Fig. 11—Curves to be used in correcting Miller's formula when the grid approaches the anode very closely. The correction term y is obtained from the equation $\frac{1}{\mu_{11}}=\frac{1}{\mu_M}-y$, where μ_M is Miller's result.

TABLE III

Amplification Factor

Nd	Δ	*T*	*Nd*	Δ	*T*
0.001	—	0.9172	0.34	0.08291	0.03025
0.002	—	0.8069	0.35	0.08741	0.02780
0.003	—	0.7424	0.36	0.09199	0.02551
0.004	0.00001	0.6966	0.37	0.09664	0.02337
0.005	0.00002	0.6611	0.38	0.10137	0.02138
0.006	0.00003	0.6321	0.39	0.1062	0.019528
0.008	0.00005	0.5863	0.40	0.1111	0.017806
0.010	0.00008	0.5508	0.41	0.1160	0.016206
0.015	0.00018	0.4863	0.42	0.1210	0.014722
0.020	0.00031	0.4406	0.43	0.1261	0.013346
0.025	0.00049	0.4052	0.44	0.1312	0.012074
0.030	0.00071	0.3762	0.45	0.1363	0.010899
0.035	0.00096	0.3518	0.46	0.1416	0.009816
0.040	0.00126	0.3307	0.47	0.1468	0.008818
0.045	0.00159	0.3123	0.48	0.1521	0.007902
0.050	0.00196	0.2956	0.49	0.1575	0.007062
0.060	0.00282	0.2670	0.50	0.1629	0.006293
0.070	0.00383	0.2430	0.51	0.1683	0.005591
0.080	0.00500	0.2223	0.52	0.1738	0.004952
0.090	0.00632	0.2042	0.53	0.1793	0.004371
0.100	0.00779	0.18817	0.54	0.1848	0.003845
0.110	0.00941	0.17381	0.55	0.1903	0.003369
0.120	0.01118	0.16084	0.56	0.1959	0.002941
0.130	0.01309	0.14905	0.57	0.2015	0.002557
0.140	0.01515	0.13828	0.58	0.2071	0.002214
0.150	0.01735	0.12839	0.59	0.2127	0.0019073
0.160	0.01969	0.11929	0.60	0.2183	0.0016354
0.170	0.02217	0.11087	0.61	0.2239	0.0013951
0.180	0.02479	0.10307	0.62	0.2296	0.0011837
0.190	0.02754	0.09582	0.63	0.2352	0.0009985
0.20	0.03042	0.08908	0.64	0.2409	0.0008371
0.21	0.03342	0.08280	0.65	0.2466	0.0006972
0.22	0.03656	0.07694	0.66	0.2522	0.0005767
0.23	0.03982	0.07147	0.67	0.2579	0.0004735
0.24	0.04319	0.06636	0.68	0.2636	0.0003856
0.25	0.04664	0.06156	0.69	0.2692	0.0003114
0.26	0.05030	0.05711	0.70	0.2749	0.0002490
0.27	0.05402	0.05292	0.71	0.2806	0.00019702
0.28	0.05785	0.04900	0.72	0.2862	0.00015392
0.29	0.06178	0.04534	0.73	0.2919	0.00011845
0.30	0.06581	0.04190	0.74	0.2975	0.00008942
0.31	0.06995	0.03869	0.75	0.3031	0.00006573
0.32	0.07418	0.03568	0.76	0.3087	0.00004643
0.33	0.07850	0.03287			

Table for evaluation of Ollendorff's formula $\mu = \frac{NS-\Delta}{T}$, where N = turns per inch, S = anode-grid distance in inches measured to the centre of a grid wire, and Δ and T are given below in terms of Nd where d is the grid-wire diameter.

It is assumed that the interelectrode spacings are not small compared to the grid pitch.

The accuracy to be obtained by simple interpolation is better than 0.25 per cent between $Nd = 0.020$ and $Nd = 0.60$ and is never less than 1 per cent between 0.001 and 0.76. By the use of second differences (see for example, "Standard Four-Figure Mathematical Tables," by Milne-Thomson and Comrie, p. 213), four-figure accuracy can be obtained.

in Fig. 11 for three different values of shadow ratio. It is unlikely that correction will be necessary for any ordinary triode.

Experimental evidence for the case in which the grid closely approaches the cathode has been given in a previous paper. The complicated equation for determining the value of μ at small currents is reduced to the approximate equation (9), which can be used in most practical cases. To assist the evaluation, a set of values of $e^{-2\pi N l_g}$ as a function of Nl_g is given in Table IV.

It will be noted that the equations determined for the cases in which the grid approaches other electrodes are based on the assumption that the grid shadow ratio is small. It is believed, however, that the terms calculated from them for the correction of Miller's formula may be used satisfactorily for correcting Ollendorff's formula even when the wires are far from thin. It is hoped that when time permits, further experimental evidence on this point may be obtained.

TABLE IV

Nl_g	$e^{-2\pi Nl_g}$	Nl_g	$e^{-2\pi Nl_g}$
0.30	0.1518	0.66	0.0158
0.32	0.1339	0.68	0.0139
0.34	0.1181	0.70	0.0123
0.36	0.1041	0.72	0.0108
0.38	0.0918	0.74	0.0096
0.40	0.0810	0.76	0.0084
0.42	0.0714	0.78	0.0074
0.44	0.0630	0.80	0.0066
0.46	0.0556	0.82	0.0058
0.48	0.0490	0.84	0.0051
0.50	0.0432	0.86	0.0045
0.52	0.0381	0.88	0.0040
0.54	0.0336	0.90	0.0035
0.56	0.0296	0.92	0.0031
0.58	0.0261	0.94	0.0027
0.60	0.0231	0.96	0.0024
0.62	0.0203	0.98	0.0021
0.64	0.0179	1.00	0.0019

Table for use with the formula for amplification factor when the distance l_g of the grid from the cathode is small compared to the pitch. In these conditions μ varies with current between limits set approximately by

$$\mu = \mu_M\left\{1 \pm 2e^{-2\pi N l_g}\left[1+\frac{l_g}{S}(\mu_M+1)\right]\right\},$$

where μ_M is the value of μ calculated for large grid-cathode spacings, N is the number of grid turns per unit length, and S is the distance of the anode from the grid plane.

Attenuation and Q Factors in Wave Guides

By A. G. CLAVIER

Laboratoire Central de Télécommunications, Paris, France

THIS PAPER begins by recalling the definitions of the attenuation and "Q" factors for coaxial lines and resonant circuits; it then discusses how these definitions can be generalized to obtain, on the one hand, the notion of attenuation per unit length for each wave structure propagated at a definite velocity in a wave guide, and on the other hand, the notion of a Q factor for a section of guide used as a resonant cavity.

The proof of a fundamental relation between these two factors, given by the author[1] as early as 1939, is then set forth. This relation allows the determination of the attenuation per unit length by measurement of the Q factor and of the energy transfer velocity, the latter being itself derived from the phase velocity. Determination of attenuation by means of these measurements is advantageous in that they may be made in the laboratory using centimeter waves and guide sections only a few meters long.

• • •

1 *Definitions and Relations for Coaxial Lines and Circuits*

Two lines of thought have been followed in setting up a theory for periodic electrical vibrations on a coaxial line. The first dates back to Kirchhoff and takes into account the intuitive existence of unit-length constants on the line (R, L, C, G). The second tries to obtain equations of the same type by deriving them from Maxwell's laws for electromagnetism. The passing from Maxwell's equations to Kirchhoff's formulation can be done easily in the ideal case where energy dissipation is assumed to be absent. In standard notation, Kirchhoff's equations then can be written:

$$-\frac{\partial u}{\partial z}=L\frac{\partial i}{\partial t}, \quad -\frac{\partial i}{\partial z}=C\frac{\partial u}{\partial t}, \tag{1}$$

and the equations derived from Maxwell's theory:

$$-\frac{\partial \varphi}{\partial z}=2\mu \log\frac{b}{a}\frac{\partial i}{\partial t}, \quad -\frac{\partial i}{\partial z}=\frac{\epsilon}{2\log\frac{b}{a}}\frac{\partial \varphi}{\partial t}. \tag{2}$$

One difference remains; φ is the scalar potential difference on which depends the work of the total electric field between two points of the cross section, one on the central conductor and the other on the outer conductor,[2] while u, according to Kirchhoff's conception, is the electrostatic difference of potential. If these two quantities are sufficiently close to be assumed equal, the two systems become identical and the values of the coefficients in (2) are those given by Kirchhoff for the L and C constants.

This transition is more difficult if the losses are not neglected, even for the coaxial line, the cylindrical structure of which is convenient for calculation and the outer conductor of which can always be assumed to be thick enough to allow any energy transmission outside the line to be ignored.

However, the very satisfactory confirmation given by experiment to the results of Kirchhoff's equations undoubtedly justifies these equations, and leads us to think that their derivation from Maxwell's equations is prevented only by difficulties in applying the approximations tacitly understood in Kirchhoff's analysis. Accordingly, we are entitled to make use of the latter without insisting on a rigorous justification. Thus, we write

$$-\frac{\partial u}{\partial z}=L\frac{\partial i}{\partial t}+Ri, \quad \text{and} \quad -\frac{\partial i}{\partial z}=C\frac{\partial u}{\partial t}+Gu. \tag{3}$$

When, as in the following discussion, only periodic sinusoidal phenomena are considered (angular frequency ω), the classical method is to relate the u and i values to the real parts of cissoidal functions (i=real part of $Ie^{j\omega t}$), and we obtain

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[1] A. G. Clavier, "Mesure des Longueurs d'Onde dans le Domaine des Ondes Décimétriques, Centimétriques et Millimétriques" (Communication to the Sixth Section, Société française des Electriciens; May, 1939), *Bulletin Société française des Electriciens*, v. 1, pp. 147–164; March, 1941.

[2] A. G. Clavier, "Conférences sur les Lois Générales Régissant la Propagation des Ondes Electromagnétiques," held at Les Laboratoires, Le Matériel Téléphonique; 1943.

$$-\frac{dU}{dz}=(R+jL\omega)U, \quad -\frac{dI}{dz}=(G+jG\omega)I, \qquad (4)$$

I and U being thus dependent on propagation equations of the same type:

$$-\frac{d^2U}{dz^2}=PU,$$

the (complex) propagation constant P being defined as

$$P^2=(R+jL\omega)(G+jC\omega). \qquad (5)$$

If the solution U is written in the form,

$$U=U_1e^{-Pz}+U_2e^{+Pz}, \qquad (6)$$

we obtain, for I

$$I=\frac{U_1}{Z_0}e^{-Pz}-\frac{U_2}{Z_0}e^{+Pz}. \qquad (7)$$

The quantity

$$Z_0=\sqrt{\frac{R+jL\omega}{G+jC\omega}} \qquad (8)$$

is the characteristic impedance of the line.

At high frequencies, the quantity $R/L\omega$ is very approximately equal to

$$\frac{\delta}{b}\,\frac{1+\frac{b}{a}}{2\log\frac{b}{a}},$$

where

δ=skin thickness$=1/\sqrt{2\pi\mu\sigma\omega}$,
b=inner radius of the outer conductor,
a=radius of the inner conductor,
σ=conductivity of the metal.

The ratio

$$1+\frac{b}{a}\Big/\log\frac{b}{a}$$

goes through a very flat maximum equal to 3.6 for $b/a=3.6$. Under such conditions, the ratio $R/L\omega$ is smaller than 0.01 for $\delta<0.005b$, and for $b=1$ centimeter if $\delta<0.005$ centimeter. The factor δ is equal to 0.004 centimeter for copper at 3 megacycles. Thus, at increased frequencies numerous experimental cases exist where $R/L\omega\ll 1$.

Similarly, for a cable whose annular space is completely filled with dielectric, $G/C\omega$ is equal to $\tan\theta$, θ being the loss angle of the dielectric. Now, for good dielectrics, at frequencies up to 600 megacycles, $\tan\theta$ is of the order of a few units times 10^{-4}.

We shall take the case where the two ratios $R/L\omega$ and $G/C\omega$ are negligible as compared with unity. It is then possible, by letting $P=\alpha+j\beta$, to obtain the approximate expressions for α and β which, together with Z_0, are often called the secondary line constants.

We find, for the attenuation per unit length α:

$$\alpha\cong\frac{R}{2}\sqrt{\frac{C}{L}}+\frac{G}{2}\sqrt{\frac{L}{C}}. \qquad (9)$$

For the phase constant $\beta(=2\pi/\lambda)$,

$$\beta\cong\omega\sqrt{LC} \qquad (10)$$

(as $\beta=\omega/v$, we obtain $v=1/\sqrt{LC}$), and for the characteristic impedance Z_0,

$$Z_0=\sqrt{\frac{L}{C}}. \qquad (11)$$

Consider now, in a case where the above approximations are valid, a line element used as a resonant circuit; a quarter-wave element for instance. The velocity of propagation being $v=1/\sqrt{LC}$, the length of this element will be

$$\frac{1}{4}\frac{1}{\sqrt{LC}}T,$$

T being the period of the oscillation used.

We shall assume the end losses (short-circuit losses and radiation losses at the input end) to be negligible as compared to the line losses. Let us find how the input impedance varies as a function of frequency f (or of wavelength, $\lambda=v/f$). This impedance is given by the classical formula

$$Z_1=Z_0\tanh(Pl) \qquad (12)$$

derived from the solution of Kirchhoff's equations.

For a line exactly a quarter-wave long, the impedance $Z_1=Z_0\tanh(\alpha+j\beta)l$ is equivalent to $Z_{\lambda/4}=\dfrac{Z_0}{\tanh\alpha l}$. It is high if $\alpha(2n+1)\lambda/4\ll 1$, and approaches zero when n becomes very high. If we consider only a number of quarter waves such that $(\alpha l)^3\ll\alpha l$, we can write as a good approximation:

$$Z_{\lambda/4}=\frac{Z_0}{\alpha l}. \qquad (13)$$

If now we move away from frequency f_0 (corresponding to the wavelength λ_0 for which the line is exactly one-quarter wave in length) by an amount δf_0, such that $\delta f_0/f_0 \ll 1$, we find

$$Z_1 = Z_{\lambda/4} \times \frac{1}{1+j\dfrac{\pi}{\alpha\lambda_0}\times\dfrac{2\delta f_0}{f_0}}. \tag{14}$$

Thus, the impedance modulus falls from 1 to $1/\sqrt{2}$ for a frequency variation δf_0 such that

$$\frac{2\delta f_0}{f_0} = \frac{\alpha\lambda_0}{\pi}. \tag{15}$$

The quantity $\pi/\alpha\lambda_0$ thus defines the selectivity of the system.

Now in antiresonant circuits with lumped constants (blocking circuits), the coefficient defined in a similar manner has been referred to in French technical literature as the "voltage magnification factor," because its value $L\omega_0/R$ is equal to the voltage magnification factor of the coil used in the antiresonant circuit. It is also agreed to call the quantity $\pi/\alpha\lambda_0$ the voltage magnification factor of the resonant line. In the American technical literature, this factor is designated by the letter Q. The International Committee for Standardization has decided, following Mr. Mesny's suggestion, to designate it by S which, in French, offers the advantage of using the initial letter common to voltage magnification and to selectivity (*Surtension* and *Sélectivité*). We shall use this notation and write

$$S = \frac{\pi}{\alpha\lambda_0}. \tag{16}$$

According to our assumptions, the α and S coefficients are related in a very simple manner in the case of the transmission line. We propose to find out what becomes of these ideas and of (16) in the more general case of wave guides. Proof of (14):

We have

$$Z_1 = Z_0 \tanh(\alpha+j\beta)l = Z_0\frac{\tanh \alpha l + \tanh j\beta l}{1+\tanh \alpha l\cdot\tanh j\beta l}$$
$$= Z_0\frac{\tanh \alpha l + j\tan\beta l}{1+j\tanh \alpha l\cdot\tan\beta l}.$$

But $\tan\beta l = \tan \omega l/v$. Let $\omega = \omega_0 + \delta\omega_0$, with $\omega_0 l/v = (2n+1)\pi/2$. Then we obtain

$$\tan\beta l = \frac{\tan\dfrac{\omega_0 l}{v}+\tan\dfrac{\delta\omega_0 l}{v}}{1-\tan\dfrac{\omega_0 l}{v}\tan\dfrac{\delta\omega_0 l}{v}} = \frac{-1}{\tan\dfrac{\delta\omega_0 l}{v}}.$$

Therefore

$$Z_1 = Z_0\frac{1+j\tanh\alpha l\cdot\tan\dfrac{\delta\omega_0 l}{v}}{\tanh\alpha l + j\tan\dfrac{\delta\omega_0 l}{v}}.$$

If we consider only cases where $\alpha l \ll 1$, and $\delta\omega_0 l/v \ll 1$, we obtain, approximately:

$$Z_1 = \frac{Z_0}{\alpha l}\times\frac{1}{1+j\dfrac{1}{\alpha l}\dfrac{\delta\omega_0 l}{v}}$$

and

$$Z_1 = Z_{\lambda/4}\frac{1}{1+j\dfrac{\pi}{\alpha\lambda_0}\dfrac{2\delta\omega_0}{\omega_0}},$$

which is (14).

2 *Application of* α *and* S *to Wave Guides*

In April 1936, Carson, Mead, and Schelkunoff,[3] of the Bell Laboratories, published an analysis of the types of propagation of electromagnetic waves in metal tubes having no central conductors, or wave guides. Concurrently, Southworth[4] and Barrow[5] have shown the agreement between theory and experimental data. In a paper presented before the Société française des Electriciens, attention was drawn to the relationships between the wave-guide theory and the theory of the coaxial cable;[6] in the latter, as in wave guides, a double infinity of elementary wave structures can be propagated, each one of them being characterized by a well-defined phase velocity. The presence of the central conductor provides an additional solution for the case of the coaxial cable; this same solution is impossible for the wave guide as it would require an infinite value for the electric and magnetic fields on the axis.[2, 5]

In following this line of thought, which is the most convenient for transmission engineers, it becomes necessary to inquire what becomes of the idea of attenuation per unit length for each wave type, and of that of Q for guide sections used as resonant elements. Then, we shall have to generalize the definitions of α and S to make them applicable to propagation and resonance phenomena in the case where the idea of primary constants (L and C) loses its intuitive value. This can be accomplished by giving an energy interpretation for these quantities. First consider (9), which gives the α coefficient for the coaxial line:

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[3] J. R. Carson, Sallie P. Mead, S. A. Schelkunoff, "Hyperfrequency Wave-guides, Mathematical Theory," *Bell System Technical Journal*, v. 15, pp. 310–333; April, 1936.

[4] G. C. Southworth, "Hyperfrequency Wave-Guides, General Considerations," *Bell System Technical Journal*, v. 15, pp. 284–309; April, 1936.

[5] W. L. Barrow, "Transmission of Electromagnetic Waves in Hollow Tubes of Metal," *Proceedings of the I.R.E.*, v. 24, pp. 1298–1328; October, 1936.

[6] A. G. Clavier, "Théorie des Câbles Diélectriques Cylindriques et Relations avec la Théorie des Câbles Co-axiaux," Communication to the fifth section, *Société française des Electriciens*, November 1937; *Bulletin Société française des Electriciens*, v. 8, pp. 355–388; April, 1938.

$$\alpha=\frac{R}{2}\sqrt{\frac{C}{L}}+\frac{G}{2}\sqrt{\frac{L}{C}}.$$

This can be written

$$\alpha=\frac{1}{2}\cdot\frac{R}{Z_0}+\frac{1}{2}GZ_0. \tag{17}$$

Consider a traveling wave (i.e., a wave for which $U_2=zero$ in (6)). We then have $i=u/Z_0$, from which $Z_0 i^2_{eff}=u^2_{eff}/Z_0$, which expresses the equipartition of the average line energies of magnetic and electric character for each period.

We see that (17) can also be written

$$\alpha=\frac{1}{2}\frac{Ri^2_{eff}+Gu^2_{eff}}{Z_0 i^2_{eff}}. \tag{18}$$

Now, $Z_0 i^2_{eff}$ is the average power for one period $\bar{P}_z$ passing through the cable cross section at abscissa z. $Ri^2_{eff}+Gi^2_{eff}$ is the average power per period lost in a unit length $\bar{q}_z$. We thus obtain for α the energy definition expressed by the formula

$$\alpha=\frac{1}{2}\cdot\frac{\bar{q}_z}{\bar{P}_z}. \tag{19}$$

Let us now turn to expression (16) for the Q factor:

$$S=\frac{\pi}{\alpha\lambda_0}. \tag{16}$$

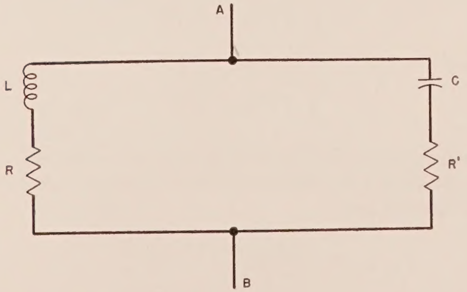

Fig. 1

In this form, an immediate generalization based on energy considerations does not appear, but we find it by reverting to the expression for the Q factor in the case of a circuit with lumped constants (Fig. 1).

For instance, in an antiresonant circuit (blocking circuit), the impedance Z, between points A and B, is given by the expression

$$\frac{1}{Z}=\frac{1}{R+jL\omega}+\frac{1}{R'-\frac{j}{C\omega}},$$

from which we derive easily

$$Z=\frac{L}{C(R+R')}\times\frac{1+\frac{R}{L\omega}CR'\omega+j\left(CR'\omega-\frac{R}{L\omega}\right)}{1+j\frac{1}{R+R'}\left(L\omega-\frac{1}{C\omega}\right)}.$$

We shall consider only those circuits operating at frequencies close to the frequency at which $L\omega-\frac{1}{C\omega}=0$ (let ω_0 be the corresponding angular frequency). We shall also assume that for ω_0 (i.e., for all the frequencies considered), $R/L\omega$ and $CR'\omega$ are very small compared with unity (inductance coils and capacitors with high Q factors). In such conditions, by letting $\omega=\omega_0+\delta\omega_0$ we have very nearly

$$Z=\frac{L}{C(R+R')}\times\frac{1}{1+j\frac{1}{R+R'}\left(\frac{1}{2}L\omega_0+\frac{1}{2}\cdot\frac{1}{C\omega_0}\right)\frac{2\delta\omega_0}{\omega_0}}. \tag{20}$$

The maximum impedance occurs at $\delta\omega_0=0$. The impedance modulus Z falls to $1/\sqrt{2}$ of its maximum value for a frequency variation $\delta\omega_0$ such that

$$\frac{\omega_0}{2\delta\omega_0}=\frac{\frac{1}{2}L\omega_0+\frac{1}{2}\cdot\frac{1}{C\omega_0}}{R+R'}$$

and we derive for the Q factor

$$S=\frac{2\pi}{T}\times\frac{\frac{1}{2}L+\frac{1}{2}C\frac{1}{C^2\omega^2}}{R+R'}. \tag{21}$$

Multiplying both numerator and denominator by i^2_{eff} (i is the current flowing through the coil), and noting that because of the high Q (very closely) $u=i/C\omega$ (u at the terminals of C), the expression for S becomes

$$S=2\pi\cdot\frac{\frac{1}{2}Li^2_{eff}+\frac{1}{2}Cu^2_{eff}}{(R+R')i^2_{eff}T}. \qquad (22)$$

S can now be defined on an energy basis as the product of 2π times the ratio of the average value of the energy stored in the circuit over one period $\overline{W}_{st}$ to the energy dissipated by virtue of the Joule effect during one period $\bar{Q}T$:

$$S=2\pi\frac{\overline{W}_{st}}{\bar{Q}T}. \qquad (23)$$

We must first determine whether this definition, applied to the coaxial line, enables us to find (16), $S=\pi/\alpha\lambda_0$.

Along the line, we have [see (6) and (7)]

$$U=U_1e^{-\alpha z}e^{-j\beta z}+U_2e^{\alpha z}e^{j\beta z}, \qquad (24)$$

$$I=\frac{U_1}{Z_0}e^{-\alpha z}e^{-j\beta z}-\frac{U_2}{Z_0}e^{\alpha z}e^{j\beta z}.$$

The line is short circuited, and no account is taken of the short-circuit impedance. We have, therefore:

$$0=U_1e^{-\alpha l}e^{-j\beta l}+U_2e^{\alpha l}e^{j\beta l},$$

that is to say,

$$U_2=-U_1e^{-2\alpha l}e^{-2j\beta l}, \qquad (25)$$

and, consequently,

$$U=U_1(e^{-\alpha z}e^{-j\beta z}-e^{-2\alpha l}e^{-2j\beta l}e^{\alpha z}e^{j\beta z}),$$

$$I=\frac{U_1}{Z_0}(e^{-\alpha z}e^{-j\beta z}+e^{-2\alpha l}e^{-2j\beta l}e^{\alpha z}e^{j\beta z}). \qquad (26)$$

Let us calculate

$$\overline{W}_{st}=\tfrac{1}{4}L\int_0^l II^*dz+\tfrac{1}{4}C\int_0^l UU^*dz \qquad (27)$$

and

$$\bar{Q}=\tfrac{1}{2}R\int_0^l II^*dz+\tfrac{1}{2}G\int_0^l UU^*dz. \qquad (28)$$

We find

$$\int_0^l UU^*dz=U_1^2\int_0^l[e^{-2\alpha z}+e^{-2\alpha(l-z)}$$
$$-2e^{-2\alpha l}\cos 2\beta(l-z)]dz, \qquad (29)$$

$$\int_0^l II^*dz=\frac{U_1^2}{Z_0^2}\int_0^l[e^{-2\alpha z}+e^{-2\alpha(l-z)}$$
$$+2e^{-2\alpha l}\cos 2\beta(l-z)]dz,$$

since Z_0 is real.

In the case of a quarter-wave line,

$$l=(2n+1)\lambda/4$$
$$\cos 2\beta(l-z)=\cos[(2n+1)\pi-2\beta z]=-\cos 2\beta z,$$

and

$$\int_0^{(2n+1)\lambda/4}\cos 2\beta z\cdot dz=\frac{1}{2\beta}(\sin 2\beta z)_0^{(2n+1)\lambda/4}=0.$$

(The same would hold for half-wave lines.) Therefore, in the present case

$$\int_0^l II^*dz=\frac{1}{Z^2_0}\int_0^l UU^*dz,$$

an equation which expresses the equipartition of the average values over one period of the magnetic and electric energies stored in the resonant line (since $Z_0^2=L/C$).

We now come back to expression (23) and obtain

$$S=\frac{2\pi}{T}\times\frac{\frac{1}{4}\cdot\frac{L}{Z_0^2}+\frac{1}{4}C}{\frac{1}{2}\cdot\frac{R}{Z_0^2}+\frac{1}{2}G},$$

which can be written

$$S=\frac{\pi}{T}\times\frac{1}{\sqrt{LC}}\times\frac{1}{\frac{R}{2}\sqrt{\frac{C}{L}}+\frac{G}{2}\sqrt{\frac{L}{C}}}$$

and gives $S=\pi/\alpha\lambda_0$.

This calculation suggests several important points which will greatly facilitate our work in the case of wave guides:

A. Within the limits in which the approximation $\beta=\omega\sqrt{LC}$ is valid, it is not necessary to take into account the attenuation α in the calculation of integrals (29).

We should have had, in this case:

$$\int_0^l UU^*dz=2U_1^2l$$

and

$$\int_0^l II^*dz=\frac{2U_1^2}{Z_0^2}l$$

and the result for S would not have been changed.

B. Following this simplified method of calculation, we find

$$S=\frac{2\pi}{T}\times\frac{\frac{1}{4}L\frac{2U_1^2}{Z_0^2}l+\frac{1}{4}C\cdot 2U_1^2l}{\frac{1}{2}R\frac{2U_1^2}{Z_0^2}l+\frac{1}{2}G\cdot 2U_1^2l},$$

which indicates that a correct result is obtained by taking the values of $\bar{W}_{st}$ and $\bar{Q}$ corresponding to a traveling wave having an amplitude $U_1\sqrt{2}$ (i.e., practically any wave) *on a unit length*. Let $\bar{W}_z$ and $\bar{q}_z$ be these values at abscissa z.

C. If we now compare the values of α and S expressed on an energy basis as functions of the quantities involved in the propagation of a traveling wave, we obtain

$$\alpha = \frac{\bar{q}_z}{2\bar{P}_z}, \quad S = 2\pi \frac{\bar{W}_z}{\bar{q}_z T}. \tag{30}$$

Hence a generalized relationship between coefficients α and S is

$$\alpha S = \pi \frac{\bar{W}_z}{\bar{P}_z T}. \tag{31}$$

We recall that in this formula $\bar{W}_z$ represents, for the traveling wave, the average value over one period of the electromagnetic energy per unit length at abscissa z. For the traveling wave, $\bar{P}_z$ represents the mean value over one period of the power going through this cross section in a unit time.

The result is that if we are dealing with the mean energy going through cross section S (Fig. 2), at abscissa z, during time dt, it can be written $\bar{W} = \bar{P}_z dt$. This mean energy is also found in a length dl, and we have $\bar{W} = \bar{W}_z dl$. The quantity $\bar{P}_z / \bar{W}_z$ or dl/dt is thus the velocity of the average energy flow along the cable. Let v_w be this velocity, and we have, finally:

$$\alpha S = \frac{\pi}{v_w T} = \frac{\pi}{\lambda_w}. \tag{32}$$

If we calculate this velocity in the case of the coaxial cable for the traveling wave, we find

$$\bar{W}_z = \tfrac{1}{4} L I I^* + \tfrac{1}{4} C U U^* = \tfrac{1}{4} L \frac{U_1^2}{Z_0^2} e^{-2\alpha z} + \tfrac{1}{4} C U_1^2 e^{-2\alpha z}$$

and

$$\bar{P}_z = \tfrac{1}{2} I U^* = \tfrac{1}{2} \cdot \frac{U_1^2}{Z_0} e^{-2\alpha z}.$$

Hence:

$$v_w = \frac{\bar{P}_z}{\bar{W}_z} = \frac{2}{Z_0\left(\frac{L}{Z_0^2} + C\right)}$$

and since

$$Z_0 = \sqrt{\frac{L}{C}},$$

$$v_w = \frac{1}{\sqrt{LC}}. \tag{33}$$

Thus, with the assumptions made, the velocity of the average energy flow for the wave traveling along the coaxial cable is equal to the phase velocity. We then have

$$\alpha S = \frac{\pi}{\frac{1}{\sqrt{LC}} T} = \frac{\pi}{\lambda_0},$$

which brings us back to (16), from which we started to obtain the correct generalization leading to (32).

We may note, again, that in the calculation of the average energy flow (33) we could have dispensed with the attenuation factor α.

3 Proof that $S = \pi/\alpha\lambda_w$ for Wave Guides

The above considerations enable us to generalize the definitions of the line coefficient α, and of the numerical coefficient S for any type of wave capable of being propagated with a definite phase velocity in a guide. Obviously, it must be assumed that the structure of the resonant sections for which S is determined is such that the two traveling waves (which move in opposite directions and interfere to produce the stationary state of electromagnetic vibrations) must have the same field configuration and consequently, the same phase velocity. This will surely take place in the case considered here, where the end wall is assumed as a perfect reflector (therefore with no losses).

The analysis of the ideal case where the guide walls do not cause any energy dissipation, leads us into classifying all possible types of simple periodic waves into two classes; the E waves and the H waves.[1, 2, 5]

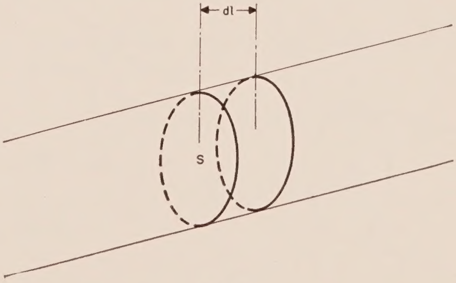

Fig. 2

Each wave type offers the phenomenon of a cutoff frequency which depends on the geometrical configuration of the wave guide; rectangular, circular, or elliptical cross section, for instance. For each wave type, and at any frequency above the cutoff frequency, the laws governing the distribution of the electric- and magnetic-field vectors are well defined. Any E or H wave has cross-section vectors $\boldsymbol{E}_s$ and $\boldsymbol{H}_s$, without whose simultaneous presence there could be no energy flow in the axial direction. But every E wave has additionally an axial field $\boldsymbol{E}_z$, which is not zero everywhere, but no $\boldsymbol{H}_z$ field, and conversely, every H wave has an $\boldsymbol{H}_z$ field with no $\boldsymbol{E}_z$ field.

For each wave type, the cross-section vectors are in phase and perpendicular to each other. In 1937 it was shown [6,7] that for any E wave, the moduli of the cross-section vectors have the relationship

$$|\boldsymbol{H}_s| = \sqrt{\frac{\epsilon}{\mu}\frac{v}{c}}|\boldsymbol{E}_s| \tag{34}$$

and similarly, for any H wave, we have

$$|\boldsymbol{E}_s| = \sqrt{\frac{\mu}{\epsilon}\frac{v}{c}}|\boldsymbol{H}_s|. \tag{35}$$

In these relationships, c is the velocity of light in the homogeneous dielectric filling the wave guide.

It will be sufficient, for the purpose of the present discussion, to relate the axial vectors to the cross-section vectors by the law of equipartition of the average values of magnetic and electric energies in any guide section; which law was used above for the coaxial wave, and which is now employed in a more general manner. Consider a guide section limited by two cross sections S_1 and S_2 (Fig. 3). The flux of the complex Poynting vector $\frac{1}{8\pi}(\boldsymbol{E}\times\boldsymbol{H}^*)$ is zero on the lateral surface, and as the $\boldsymbol{E}_s$ and $\boldsymbol{H}_s$ cross-section vectors are in phase, it is real through the sections S_1 and S_2.

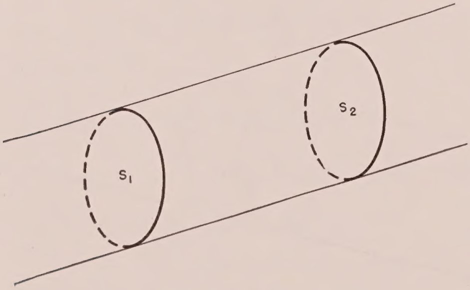

Fig. 3

The result is, according to Green's theorem, that the volume integral

$$\int_v \frac{1}{8\pi}\,\mathrm{div}\,\boldsymbol{E}\times\boldsymbol{H}^* d\tau,$$

extended to the volume of the guide limited by cross sections S_1 and S_2, must be a real quantity.

Now we have

$$\mathrm{div}\,\boldsymbol{E}\times\boldsymbol{H}^* = \boldsymbol{H}^*\,\overline{\mathrm{curl}\,\boldsymbol{E}} - \boldsymbol{E}\,\overline{\mathrm{curl}\,\boldsymbol{H}^*}$$

and from Maxwell's equations,

$$\overline{\mathrm{curl}\,\boldsymbol{E}} = -j\omega\mu\boldsymbol{H}, \quad \overline{\mathrm{curl}\,\boldsymbol{H}} = j\omega\epsilon\boldsymbol{E},$$

we obtain

$$\int_v \frac{1}{8\pi} dw\boldsymbol{E}\times\boldsymbol{H}^* d\tau = 2j\omega\int_v\left(\frac{\epsilon}{8\pi}\frac{\boldsymbol{EE}^*}{2} - \frac{\mu}{8\pi}\frac{\boldsymbol{HH}^*}{2}\right)d\tau$$
$$= 2j\omega(W_e - W_m).$$

As this expression must be real, the right-hand member must be zero; we do have an equipartition of the average values of the magnetic and electric energies.

We are now able, for any wave type, to calculate the velocity v_w of average energy flow as a function of the phase velocity. Consider, for example, an E wave (the same reasoning would hold for an H wave). We have

$$v_w = \frac{\bar{P}_z}{\bar{W}_z},$$

but

$$\bar{P}_z = \frac{1}{8\pi}\int_s \boldsymbol{E}\times\boldsymbol{H}^* ds = \frac{1}{8\pi}\int_s E_s H_s^* ds$$

and

$$\bar{W}_z = \frac{1}{16\pi}\int_v (\epsilon EE^* + \mu HH^*) d\tau,$$

volume v being constituted by a segment of cross section S and unit length.

As in the case of the coaxial cable, we shall assume that losses are sufficiently low not to disturb appreciably the laws governing the distribution of fields found in the ideal case where there

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[7] L. Brillouin, "Theoretical Study of Dielectric Cables," *Electrical Communication*, v. 16, pp. 350–372; April, 1938.

are no losses. In these conditions, E and H are dependent on z through exponential $e^{j\omega(t-z/v)}$ only; the z variable disappears in the products to be integrated, and since we integrate on a unit length, we can write

$$\bar{P}_z=\frac{1}{8\pi}\int_v E_s\cdot H_s^* d\tau$$

and because of relation (34),

$$\bar{P}_z=\frac{1}{8\pi}\sqrt{\frac{\epsilon}{\mu}}\,\frac{v}{c}\int_v E_sE_s^* d\tau.$$

Also, because of the energy equipartition,

$$\bar{W}_z=\frac{1}{16\pi}2\int_v \mu H_sH_s^* d\tau=\frac{1}{8\pi}\epsilon\frac{v^2}{c^2}\int_v E_sE_s^* d\tau$$

and since $c=1/\sqrt{\epsilon\mu}$, we obtain

$$v_w=\frac{\bar{P}_z}{\bar{W}_z}=\frac{c^2}{v}. \qquad (36)$$

Thus, for waves propagated in guides, the velocity of the mean energy transfer is equal to the group velocity. The above proof is taken from my lectures[8] before the radio division of the Ecole Supérieure d'Electricité. It is particularly simple. Other means of proving this property will be found in L. de Broglie's book,[9] and in my 1943 lectures[2] to the engineers of Les Laboratoires, Le Matériel Téléphonique.

Starting from (31),

$$\alpha S=\pi\frac{\bar{W}_z}{\bar{P}_z T}.$$

This relationship leads to (32):

$$\alpha S=\frac{\pi}{\lambda_w}=\frac{\pi}{\frac{c^2}{v^2}\lambda_p}, \qquad (37)$$

which is valid for wave guides. The factor λ_p is the phase wavelength along the guide, a quantity which is directly measurable by means of stationary waves and when known, together with the frequency, enables $v(=\lambda_p f)$ to be determined.

The relationship $S=\pi/\alpha\lambda_w$ was mentioned as early as 1939 in a communication[1] to the Société française des Electriciens.

4 Effect of Energy Dissipation in the End Walls

The above considerations do not take into account the energy dissipation in the end walls. To explain the effect of this loss on coefficient S, we could again start from the analysis of the propagation phenomena in a resonant coaxial line terminated by a resistance r (Fig. 4). In case the latter, however, is small as compared with the characteristic impedance Z_0, we then have instead of $S=\pi/\alpha\lambda_0$, the expression

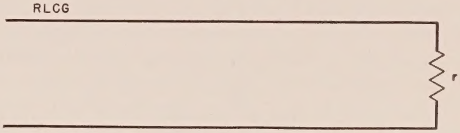

Fig. 4

$$S'=\frac{\pi}{\left(\alpha+\frac{r}{Z_0 l}\right)\lambda_0}. \qquad (38)$$

But this expression can be found more quickly by starting with the energy expression (23) of the Q factor,

$$S=2\pi\frac{\bar{W}_{st}}{\bar{Q}T},$$

and by assuming that resistance r is small enough not to disturb appreciably the law of current distribution along the line. We thus obtain

$$S'=2\pi\frac{\bar{W}_{st}}{(\bar{Q}+\bar{q})T}, \qquad (39)$$

where $\bar{q}$ is the mean power dissipated in resistance r. But

$$\bar{W}_{st}=\tfrac{1}{4}L\frac{2U_1^2}{Z_0^2}l+\tfrac{1}{4}C\cdot 2U_1^2 l,$$

and

$$\bar{Q}=\tfrac{1}{2}R\frac{2U_1^2}{Z_0^2}l+\tfrac{1}{2}G\cdot 2U_1^2 l,$$

now:

$$\bar{q}=\tfrac{1}{2}r\left(\frac{2U_1}{Z_0}\right)^2=r\frac{2U_1^2}{Z_0^2}.$$

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[8] A. G. Clavier, "Les Ondes Electromagnétiques Centimétriques," Lectures at the Ecole Supérieure d'Electricité; 1942.

[9] L. de Broglie, "Problèmes de Propagations Guidées des Ondes Electromagnétiques," Gauthier-Villars (Paris); 1941.

After a few transformations, we again obtain expression (38):

$$S'=\frac{\pi}{\alpha+\frac{r}{Z_0 l}\lambda_0}.$$

Before mentioning how (37) leads to a particularly interesting measurement of the attenuation factor per unit length (α) in guided waves, it may not be without interest to make a few remarks on the energy definitions of the quantities involved in the discussion. Let us recall them:

$$\alpha=\tfrac{1}{2}\cdot\frac{\bar{q}_z}{\bar{P}_z}, \tag{19}$$

$$v_w=\frac{\bar{P}_z}{\bar{W}_z}, \tag{32}$$

$$S=2\pi\frac{\bar{W}_z}{\bar{q}_z T}=\frac{\pi}{\alpha v_w T} \tag{30}$$

and

$$S'=2\pi\frac{\bar{W}_{st}}{(\bar{Q}+\bar{q})T}. \tag{39}$$

In these definitions, (39) alone does not call for the idea of a transmission line, and applies even when the notion of transmission velocity does not appear, i.e., in the case of circuits with lumped constants. We started from such a case to find (39).

In guided waves, particularly, the case of resonance at the cutoff frequency occurs: v_w becomes zero, α becomes infinite, but both experiment and theory show that $\pi/\alpha v_w T$ tends towards a definite limit, as could be expected from (39).

Consider now any resonant element in which the mean value of electromagnetic energy is $\bar{W}_{st}$; let us allow this energy to be gradually dissipated. After a time dt, it will have decreased by

$$d\bar{W}_{st}=-(\bar{Q}+\bar{q})dt.$$

Therefore, the law of dissipation is

$$\bar{W}_{st}=W_0 e^{-t\frac{(\bar{Q}+\bar{q})}{\bar{W}_{st}}},$$

or in other words, $S=2\pi\frac{\theta}{2}\cdot\frac{1}{T}$, where we call θ the time constant of the system (referred to one of the factors whose product defines the energy). We thus have, in a general manner:

$$S=\pi\frac{\theta}{T}, \tag{40}$$

which is the product of π by the number of periods contained in the time constant. If (30) applies, ($\bar{q}\ll\bar{Q}$), we can then note that $1/\alpha$ is the distance taken by one of the factors of energy to decrease in the ratio $1/e$. Let $1/\alpha=\Lambda$ be this "length constant." We then have

$$S=\pi\frac{\Lambda}{\lambda_w}. \tag{41}$$

5 Determination of α by Measurement of Q and Velocity of Energy Transfer

Formulas (30) and (39) suggest the determination of coefficient α by measuring S' and v_w (or λ_w).

We have

$$S'=\frac{2\pi\bar{W}_{st}}{\bar{Q}T}\times\frac{1}{1+\frac{\bar{q}}{\bar{Q}}}=\frac{\pi}{\alpha\lambda_w}\times\frac{1}{1+\frac{\bar{q}}{\bar{Q}}}.$$

All that is required is to operate a line with a number of half-waves sufficient for the ratio of the end losses to the line losses to be small as compared to unity. This will be the case when the value of S' is found experimentally to remain constant from one resonance to the next; the phase wavelength, and from it $\lambda_w(=\lambda_0^2/\lambda_p)$, is determined during the same measurements. This method, which makes it possible to work on short lengths of cable, is particularly convenient for studying the effect of the nature and structure of the walls; it has already yielded interesting results which will be published later. The measurement of the Q factor is effected by means of a "Q-meter," a description[10] of which was presented before the Société française des Electriciens in 1942.

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[10] A. G. Clavier, V. Altovsky and R. Zwobada, "Emploi des Volumes Résonants dans la Technique des Ondes Centimétriques" (Communication presented to the Fifth Section, Société française des Electriciens; April 28, 1942), Notice XV, Compagnie Générale de Constructions Téléphoniques—Le Matériel Téléphonique.

Piezoelectric Substances

By M. BRUZAU

Laboratoire Central de Télécommunications, Paris, France

Editor's Note: This paper was published originally as a monograph "Les Substances Piézoélectriques Synthétiques" in French in 1940. It is reprinted in Electrical Communication inasmuch as it is felt that the classification of piezoelectric crystals still is of considerable interest. As stated in the concluding paragraph, the vast field of artificial piezoelectric bodies has as yet hardly been touched; actually, since the article was prepared, substances like dihydrogen ammonium and potassium phosphate have been used.

AT THE TIME of their discovery of piezoelectricity, in 1880, J. and P. Curie listed about 10 substances having this property. Since then the number of known piezoelectric substances has increased considerably, thanks to a very sensitive method of observation indicated by Giebe and Scheibe.

These substances, more than 200 in number, belong to crystallographic classes having no center of symmetry. Some are natural mineral substances, others may be obtained artificially. Among the latter, Rochelle salt has been the most widely studied. Its piezoelectric modulus d_{14} is the highest known (4000 times larger than the d_{11} modulus of quartz); consequently, its use has been suggested for filters, microphones, pickup devices, loudspeakers, relays, and light valves. The other artificial piezoelectric substances are little known (the moduli have been measured for only 10 of them) and, until now, they have received no practical application.

• • •

1 Historical

The discovery of piezoelectricity was made by J. and P. Curie, who published their first papers[1] in 1880. They showed that under the effect of a compressional or expansional stress, certain crystals become electrified. They looked for the cause of this new phenomenon and ascribed it to the absence of a center of symmetry in the crystal structure.

Their experiments were made with tourmaline, quartz, blende, sodium chlorate, boracite, calamine, topaz, right-handed tartaric acid, Rochelle salt, and sugar. The two scientists remarked:

> It can be foreseen that there exist many other piezoelectric substances amongst artificial crystallized substances. Substances which can act on polarized light, for instance, yield crystals in which some diameters have dissimilar extremities.

In spite of this remark, the number of known piezoelectric substances remained small for a long time. We may quote the discovery by Hankel and Lindenberg[2] in 1893 of the piezoelectricity of resorcin. In 1911, Van der Veen[3] indicated a few new piezoelectric substances, particularly benzil, sodium-lithium molybdate, barium antimoniltartrate, patchouli camphor, and rubidium tartrate. But it was not until 1924 that Lucas[4] verified, on more than 30 optically active crystals belonging to various chemical groups, that there seems to be no exception to the rule that all active substances give piezoelectric crystals. He also observed that isodulcite, quebrachite, potassium iridotrioxalate, and basic cinchonine sulfate are noticeably more piezoelectric than quartz.

2 Static Test for Piezoelectricity

In the first experiments, a static method was used: A mechanical straining of the crystal caused the appearance of electric charges which were measured with an electrometer. This method, which allows a quantitative evaluation of piezoelectricity, has the drawback of requiring fine specimens, the crystallographic axes of which can easily be determined.

3 Dynamic Test for Piezoelectricity

In 1925, Giebe and Scheibe[5] suggested a dynamic method applicable to small fragments of crystalline substance, with irregular shapes. It is based on the production of elastic vibrations in piezoelectric crystals under the action of a high-frequency alternating electric field.[6]

A large number of crystal fragments, 0.1 to 5 millimeters in size, which constitute resonators with different elastic frequencies, are arranged between the two plates of a capacitor $C1$ of Fig. 1. This capacitor is in parallel with a variable capacitor $C2$ in an oscillating circuit. The inductor and capacitors are such that, by means of the variable capacitor $C2$, a range of frequencies from 10^6 to 10^7 cycles ($\lambda = 300$ to 30 meters) can be covered.

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[1] Numbered references will be found in the bibliography on page 459.

For oscillation frequencies in the vicinity of the natural frequencies of the small crystal elements, mechanical oscillations of large amplitude can be produced in the crystals. These oscillations, by virtue of the piezoelectric properties of the substance, react on the exciting electric field and affect the frequency of the main oscillator.

The small crystal element in which resonance is set up behaves like an electric circuit coupled to the main circuit. When the variable capacitance of this main circuit is changed slowly, a frequency and amplitude discontinuity occurs near the natural frequency of the crystal. The crystal tends to hold the oscillator frequency in step until, with further detuning of the oscillator circuit, the natural oscillation of the crystal decreases in amplitude and the main oscillator suddenly resumes the frequency it would have had in the absence of the piezoelectric crystal. If the system is coupled to a receiver comprising a detector D, an amplifier A, and a telephone receiver T, a characteristic click is heard when the discontinuity occurs.

To excite a single crystal, the electrical axis of the crystal must be appropriately oriented in the field. By using a large number of crystal fragments, whose axes have random directions in space, there is always a number of crystals in the right position with respect to the electric field. Furthermore, with a large number of small resonators, the absorption band extends over a broad frequency range, while with a single crystal the natural frequency is not always easy to find and response of the telephone receiver may not be perceptible.

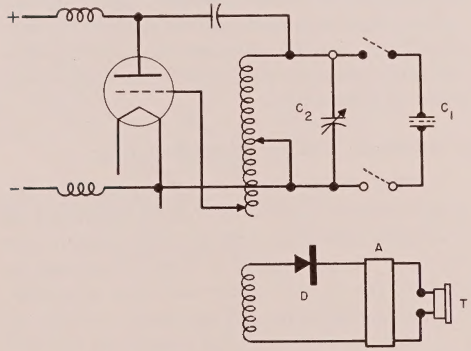


Fig. 1—Circuit for dynamic testing for piezoelectric activity of small irregular particles which are placed between the plates of C1 and subjected to an electric field which may be varied continuously between frequencies of 1 and 10 megacycles by adjustment of C2. Piezoelectric activity produces clicks in the telephone receiver of the detector-amplifier circuit.

This method is highly sensitive: Giebe and Scheibe obtained a very sharp sound in the telephone receiver by using a sample of arsenical blende of 0.03 gram divided into 40 fragments of 1 to 0.1 millimeter. It only allows, it is true, a qualitative determination; indeed, the experimental conditions being very difficult to duplicate (same cut for the fragments, same positions, etc.), no certain conclusion can be drawn from the responses in the telephone receiver caused by two different substances.

It must be noted, however, that Bergmann[7] in 1935 and Engl and Leventer[8] in 1937 showed that it was possible to use dynamic methods for quantitative measurements of piezoelectricity.

It has been observed that hygroscopic conditions and the purity of the sample may influence the production of high-frequency oscillations. Similarly, certain soft substances like camphor, give a clear result only when the temperature is lowered to −20 degrees centigrade as the oscillations are damped in the crystal at room temperature. Finally, conducting substances or those liable to become conducting through alteration cannot be studied by this method. The Giebe and Scheibe method has made it possible to extend the investigation to a large number of substances.

4 Classification

It has been stated that J. and P. Curie ascribed the piezoelectric property to an absence of symmetry in the crystal structure:

> In a centered medium, a directional phenomenon can exist only if the cause itself is directional. A vector can excite therein another vector; for instance, a field (electric or magnetic) can create a polarization (electric or magnetic). A tensor can create therein another tensor, for instance a tension or a pressure can cause an expansion or a contraction. But a tensor cannot excite a vector. The dissymmetry, not residing in the agent, must reside in the medium. Therefore the only media in which piezoelectricity is possible are without a center.[9]

4.1 HEMIHEDRISM

Crystalline substances are divided into 32 classes, 21 of which have no center of symmetry. The noncentered crystals may exhibit two kinds of hemihedrism.

A. Hemimorphous hemihedrism, antimorphous or antihemihedrism (dichosymmetrical hemiaxism of Bravais, hemihedrism with inclined faces): the number of axes is decreased by one-half, the planes of symmetry corresponding to the vanished axes are left. Conjugate types can be superposed; the crystal has different pointings at the two ends of the remaining principal axis.

B. Holoaxial hemihedrism, not capable of superposition, or enantiomorphous (plagihedral) hemihedrism: all axes of symmetry are left but there is no center and, therefore, no planes of symmetry. The two conjugate forms (right- and left-hand) cannot be superposed, generally; they are enantiomorphous.

Pasteur[10] observed that these crystals possess a rotatory power. He called attention to the fact that the cause of hemihedrism can have two distinct origins:

> It may reside in the chemical molecule itself and be carried over to any combination of these molecules; this is the case for tartaric acid. On the other hand, the shape dissymmetry may be only a consequence of the mode of aggregation of the molecules inside the crystal and this is probably the case for quartz.

5 Later Detection of Piezoelectric Substances

It must be noted that in many cases, hemihedrism of crystals will be revealed by a crystallographic investigation only in the case of a generally oblique form. On orthoparallel forms (i.e., parallel to an even-order axis), the suppression of the center does not cause any change in the external aspect of the crystal. In such a case, physical properties like pyroelectricity, piezoelectricity, and rotatory power, may sometimes allow the detection of the true symmetry of the crystal.[4, 11, 12] Thus, two methods are possible for the detection of piezoelectric substances:

A. An examination of the crystals described as belonging to one of the classes with no center of symmetry.

B. An examination of the crystals having rotatory power.

Between 1925 and 1932 about 85 new piezoelectric substances were found in this manner. Giebe and Scheibe,[5] Elings and Terpstra,[13] Hettich and Schleede,[14] Schneider,[15] and Greenwood and Tomboulian,[16] contributed to these investigations. Since 1932 the number of piezoelectric substances has been further increased by the works of Hettich and Steinmetz,[11] Greenwood,[17] Busch and Scherer,[18] Theilacher,[19] Neuhaus,[20] and Bahr and Engl.[21]

The investigators who followed the first method knew the crystallographic class of the substances investigated and indicated it. Generally those who followed the second method did not indicate the crystalline system of the substances examined. Finally, a number of scientists, especially Hettich and Steinmetz, undertook more particularly the use of piezoelectricity for determining the structures of the molecules and the nature of the linkings between their atoms or atom groups. They examined a large number of substances, both mineral and organic, regardless of their crystallographic class, which is not given in the paper.

6 Tabulation of Piezoelectric Substances

In Tables I–VII, the known piezoelectric substances have been grouped according to their crystalline systems (which has the advantage of giving an indication of the symmetry of the crystal and of the number and nature of its moduli[22]). In setting up these tables, Groth's treatise[23] was consulted when the papers gave no information as to the crystallographic class. It will be noted that a number of piezoelectric substances (especially among the organic products examined by Hettich and Steinmetz) are described as belonging to classes having a center of symmetry; their crystallographic study should be checked. Finally Table VIII includes the substances for which we could not find the crystalline system. We indicate, vertically, in the first left-hand column of the tables, the nature of hemihedrism, the class according to Groth, Nigli's notation, and the remaining symmetry elements. The numbers in the last column refer to the bibliography. The minus sign (−) means that the substance does not exhibit piezoelectricity, while the plus sign (+) or

TABLE I

Cubic System

Holohedral Form—$3\Lambda^4\ 4\Lambda^3\ 6L^2\ C\ 3\Pi\ 6P$

	Name	Symbol	Reference
Enantimorphous Hemihedrism Class: Icositetrahedral Pentagon 0 $3\Lambda_4\ 4\Lambda^3\ 6L^2$	Ammonium Chloride	NH_4Cl	+21, −14
	Hexamethylene Tetramine	$(CH_2)_6N_4$	13, 14
Hemimorphous Hemihedrism Class: Hexakistetrahedral T_d $3\Lambda^2\ 4\Lambda^3\ 6P$	Zinc Sulfide (Blende)	ZnS	I, 13, 14
	Boracite	$B_{16}O_{30}Cl_2Mg_7$	I, 13
	Diamond	C	−13
	Cuprous Chloride	$CuCl$	14
	Ammonium Fluoride	NH_4F	11
Tetartohedrism Class: Tetrahedral Dodecahedral Pentagon T $3\Lambda^2\ 4\Lambda^3$	Sodium Chlorate	$NaClO_3$	1, 5
	Sodium Thioantimoniate	$Na_3SbS_4 \cdot 9H_2O$	5
	Barium Nitrate	$Ba(NO_3)_2$	−5
	Strontium Nitrate	$Sr(NO_3)_2$	−5
	Lead Nitrate	$Pb(NO_3)_2$	−5
	Sodium Bromate	$NaBrO_3$	13, 14
No class indicated	Helvin		1
	Potassium Cadmium Nitrite	$KCd(NO_2)_3$	11
	Zinc Potassium Cyanide	$KCN \cdot Zn_2(CN)_2$	11

TABLE II

Senary or Hexagonal System

Holohedral Form—$\Lambda^6\ 3L^2\ 3L'^2\ C\ \Pi\ 3P\ 3P'$

	Name	Symbol	Reference
Enantiomorphous Hemihedrism Class: Hexagonal Trapezohedral D_6 $\Lambda_6\ 3L^2\ 3L'^2$	This symmetry is fairly rare; it would be that of quartz above 575 degrees centigrade		
Hemimorphous Hemihedrism (antihemihedrism with a senary axis) Class: Dihexagonal Pyramidal C_{6v} $\Lambda_6\ 3P\ 3P'$	Zincite	ZnO	−13 conductor 14
	Ice	H_2O	−14
	Silver Iodide	AgI	+11, −14
Tetartohedrism (with senary axis) Class: Hexagonal Pyramidal C_6 Λ_6	Potassium Lithium Sulfate	$KLiSO_4$	13
	Nepheline	$Na_8Al_8SiO_{34}$	13
	Lead Antimonyltartrate	$Pb(SbO)_2(C_4H_4O_6)_2$	16
Sphenoidal Hemihedrism Class: Ditrigonal Bipyramidal D_{3h} $\Lambda^3\ 3L^2\Pi\ 3P'$	Benitoite	$BaSi_3TiO_9$	−13, 14
Sphenoidal Tetartohedrism Class: Trigonal Bipyramidal C_{3h} $\Lambda^3\Pi$	Very rare symmetry, no known piezoelectric substance		
No class indicated	Barium Nitrite	$Ba(NO_2)_2 \cdot H_2O$	11
	Menthol	$C_{10}H_{20}O$	11
	Borneo Camphor (Borneol)	$C_{10}H_{17}OH$	11

TABLE III

TRIGONAL, TERNARY, OR RHOMBOHEDRAL SYSTEM

Holohedral Form—$\Lambda^3\ 3L^2\ C\ 3P$

	Name	Symbol	Reference
Enantiomorphous Hemihedrism Class: Trigonal Trapezohedral D_3 $\Lambda^3\ 3L^2$	Quartz	SiO_2	1, 5
	Benzil	$C_{14}H_{10}O_2$	5
	Cinnabar	HgS	13
	Guaiacol	$C_6H_4(OCH_3)^1(OH)^2$	11
	Rubidium Tartrate	$Rb_2C_4H_4O_6$	15
	Potassium Rhodoxalate	$K_2[Rh(C_2O_4)_3]\cdot H_2O$	4
	Potassium Iridotrioxalate	$K_3[Ir(C_2O_4)_3]\cdot 2H_2O$	4
Hemimorphous Hemihedrism Class: Ditrigonal Pyramidal C_{3v} $\Lambda_3\ 3P$	Ruby Silver Ore (Proustite)	Ag_3AsS_3	5
	Dark Red Silver Ore (Pyrargirite)	Ag_3SbS_3	−5, +11
	Tourmaline		1, 5
	Potassium Bromate	$KBrO_3$	13, 14
	Carborundum	SiC	−13
Tetartohedrism Class: Trigonal Pyramidal C_3 Λ_3	Sodium Periodate	$NaIO_4$	−11
No class indicated	Acetamide	CH_3CONH_2	15

TABLE IV

TETRAGONAL OR QUATERNARY SYSTEM

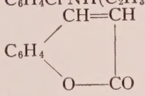
Holohedral Form—$\Lambda^4\ 2L^2\ 2L'^2\ C\ \Pi\ 2P\ 2P'$

	Name	Symbol	Reference
Enantiomorphous Hemihedrism Class: Tetragonal Trapezohedral D_4 $\Lambda^4\ 2L^2\ 2L'^2$	Nickel Sulfate	$NiSO_4\cdot 6H_2O$	5
	Strychnine Sulfate	$(C_{21}H_{22}O_2N_2)_2H_2SO_4\cdot 6H_2O$	13
	Guanidine	$HN{=}C(NH_2)_2H_2CO_3$	11
	β Methylglucoside	$C_6H_{11}O_5OCH_3$	inappreciable 11
Hemimorphous Hemihedrism (with quaternary axis) Class: Ditetragonal Pyramidal C_{4v} $\Lambda^4\ 2P\ 2P'$	Pentaerytritol	$C_5H_{12}O_4$	5
Tetartohedrism (with quaternary axis) C_4 Λ^4	Lead Molybdate	$PbMoO_4$	−13, −14
	Zinc Fluoberylate	$ZnBeF_4\cdot 7H_2O$	−17
			−17
Sphenoidal Hemihedrism Class: Tetragonal Scalenohedral V_d $\Lambda^2\ 2L^2\ 2P'$	Chalcopyrite	$CuFeS_2$	16
	Primary Ammonium Orthophosphate	$(NH_4)H_2PO_4$	5
	Primary Potassium Orthophosphate	KH_2PO_4	13
	Primary Ammonium Orthoarsenate	$(NH_4)H_2AsO_4$	13
	Primary Potassium Orthoarsenate	KH_2AsO_4	13
	Tetramethyl Ammonium Iodide	$N(CH_3)_4I$	−24
	Tetraethyl Ammonium Iodide	$N(C_2H_5)_4I$	5
	Mercuric Cyanide	$Hg(CN)_2$	13, 14
	Pentaerythritol Tetranitrate	$C(CH_2ONO_2)_4$	14
	Urea	CH_4N_2O	5
Sphenoidal Tetartohedrism Class: Tetragonal Disphenohedral S_4 Λ^2	No example		
No class indicated	Glucinium Calcium Sodium Fluosilicate	$Si_3O_{10}FBe_2Ca_2Na$	13
	Dibenzylcamphor	$C_{24}H_{28}O$	4
	Potassium Bromomethane Trisulfonate	$CBr(SO_3K)_3\cdot H_2O$	50

TABLE V

TERBINARY, ORTHORHOMBIC, OR RHOMBIC SYSTEM

Holohedral Form—$L^2\ L'^2\ L''^2\ C\ P\ P'\ P''$

	Name	Symbol	Reference
Enantimorphous Hemihedrism Class: Bisphenoidal V $L^2\ L'^2\ L''^2$ (This class is mentioned by Giebe and Scheibe as being strongly piezoelectric)	Magnesium Sulfate	$MgSO_4 \cdot 7H_2O$	5
	Zinc Sulfate	$ZnSO_4 \cdot 7H_2O$	5
	Nickel Sulfate	$NiSO_4 \cdot 7H_2O$	14
	Silver Nitrate	$AgNO_3$	−14, 17
	Strontium Formiate	$Sr(HCO_2)_2 \cdot 2H_2O$	5, 13
	Barium Formiate	$Ba(HCO_2)_2$	14
	Lead Formiate	$Pb(HCO_2)_2$	14
	Right-Hand Potassium Bitartrate	$KHC_4H_4O_6$	5, 13
	Right-Hand Ammonium Bitartrate	$(NH_4)HC_4H_4O_6$	5
	Rubidium Bitartrate	$RbHC_4H_4O_6$	11
	Potassium Sodium Tartrate (Rochelle salt)	$NaKC_4H_4O_6 \cdot 4H_2O$	1, 5
	Potassium Antimonyltartrate (emetic)	$K(SbO)C_4H_4O_6 \cdot H_2O$	14
	Barium Antimonyltartrate	$Ba[(SbO)C_4H_4O_6]_2 \cdot H_2O$	3
	Potassium Antimonyltartrate and Sodium Chloride	$4[K(SbO)C_4H_4O_6] \cdot NaCl$	16
	Monohydrated Ammonium Oxalate	$(NH_4)_2C_2O_4 \cdot H_2O$	11
	Potassium Ammonium Oxalate	$K(NH_4)C_2O_4$	11
	Sodium Citrate	$Na_3C_6H_5O_7 \cdot 5H_2O$	−14, 17
	Sulfur (from CS_2)	S	−14
	Iodic Acid	HIO_3	14
	α Aminoglutaric Acid or α Glutamic Acid	$HO_2C \cdot H_2C \cdot CH(NH_2)CO_2H$	13
	Asparagine	$C_4H_8N_2O_3 \cdot H_2O$	4, 5
	Hippuric Acid	$C_6H_5CONHCH_2 \cdot CO_2H$	13, 14
	Benzophenone	$C_6H_5 \cdot CO \cdot C_6H_5$	+13, 11, −14
	Santonin	$C_{15}H_{18}O_3$	17
	Sodium Santoninate	$NaC_{15}H_{19}O_4 \cdot 3,5H_2O$	13
	Salicine	$C_{13}H_{18}O_7$	14
	Fructose	$C_6H_{12}O_6$	14
	α Methylglucoside	$C_6H_{11}O_5OCH_3$	11
	Strychnine	$C_{21}H_{22}O_2N_2$	13, 14
	Cinchonidine	$C_{19}H_{22}ON_2$	13
	4-6 Diiodo 1-3 Dinitrobenzene	$C_6H_2I_2^{(4-6)}(NO_2)_2^{(1-3)}$	11
	Ephedrine Hydrochlorate	$C_{10}H_{15}ONHCl$	11
	Cobalt Complex	$[Co(NH_3)_4(NO_2)_2]NO_3$	11
	Phthalic Anhydride	$C_8H_4O_3$	15
	Benzal Camphor	$C_{17}H_{20}O$	4
	Anisal Camphor	$C_{18}H_{22}O_2$	4
	Benzyl Camphor	$C_{17}H_{22}O$	4
	Anisyl Camphor	$C_{18}H_{24}O_2$	4
	Bromobenzyl Camphor	$BrOC_{17}H_{21}$	4
	α Methyl Bicamphorate	$C_{11}H_{18}O_4$	4
	Ergothioneine Hydrochlorate	$C_9H_{15}O_2N_3S \cdot HCl \cdot 2H_2O$	4
Hemimorphous Hemihedrism Class: Pyramidal C_{2v} $L^2\ P'\ P''$	Calamine	$Zn_2H_2SiO_5$	1, 13
	Aragonite	$CaCO_3$	−14
	Strontium Chlorate	$Sr(ClO_3)_2$	16
	Resorcine	$C_6H_6O_2$	5
	Picric Acid	$C_6H_2(OH)(NO_2)_3^{(2-4-6)}$	−11, 17
	m. Chloronitrobenzene	$C_6H_4Cl^1(NO_2)^3$	11
	m. Bromonitrobenzene	$C_6H_4Br^1(NO_2)^3$	11
	m. Phenylene Diamine	$C_6H_4(NH_2)_2^{1-3}$	11
	Triphenylmethane	$CH(C_6H_5)_3$	5
	p. Chloracetanilide	$C_6H_4Cl^1NH(C_2H_3O)^4$	16
	Coumarin		11
Holohedrism* Class: Bipyramidal V_h $L^2\ L'^2\ L''^2\ C\ PP'\ P''$	Rubidium Nitrate	$RbNO_3$	11
	m. Dinitrobenzene	$C_6H_4(NO_2)_2^{1-3}$	11
	1-3-5 Tribromoaniline	$C_6H_2(NH_2)_2Br_3^{1-3-5}$	11
	m. Nitroaniline	$C_6H_4(NO_2)^1(NH_2)^3$	11
	1-3 Diamine 6 Methyl Benzene	$C_6H_3(NH_2)_2^{1-3}(CH_3)^6$	11
	2-6 Dinitrophenol	$C_6H_3(OH)^1(NO_2)_2^{(2-6)}$	11
	2-6 Dinitrobenzoic Acid	$C_6H_3(CO_2H)^1(NO_2)_2^{(2-6)}$	11
	1-8 Dinitronapthalene	$C_{10}H_6(NO_2)_2^{(1-8)}$	11
	p. Phenol Potassium Sulfonate	$C_6H_4OHKSO_3^4$	11
	Benzoyl Peroxide	$(C_6H_5COO)_2$	11
	Anthranilic Acid	$C_6H_4(CO_2H)^1(NH_2)^2$	11

* Although described by Groth as holohedral, these crystals, being piezoelectric, can have no center of symmetry.

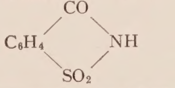
TABLE V—*Continued*

	Name	Symbol	Reference
No class indicated	Silver Iodate	$AgIO_3$	11
	Scapolite		-11, $+17$
	Methylurea	$CONH_2NH(CH_3)$	17
	Diphenylurea, Symmetric	$CO(NHC_6H_5)_2$	11
	Diphenylurea, Dissymetric	$H_2NCO—N(C_6H_5)_2$	11
	Triphenyl Guanidine	$C_6H_5—N{=}C(NHC_6H_5)_2$	11
	Diglycolic Amide Acid	$CO_2H—CH_2O—CH_2CONH_2$	11
	Succinic Anhydride	$C_4H_4O_3$	15
	o. Iodonitrobenzene	$C_6H_4I^1(NO_2)^2$	11
	2-6 Dinitrotoluene	$C_6H_3(NO_2)_2{}^{(2-6)}CH_3{}^4$	11
	Mixture, m. Bromonitrobenzene and	$C_6H_4Br^1(NO_2)^3$	11
	m. Iodonitrobenzene	$+2[C_6H_4I^1(NO_2)^3]$	11
	Dihalogenated Anilines	$C_6H_3(NH_2)^6Cl_2{}^{1-3}$	11
		$C_6H_3(NH_2)^6Br_2{}^{1-3}$	11
		$C_6H_3(NH_2)^6I_2{}^{1-3}$	11
	m. Hydroxybenzoic Acid	$C_6H_4(CO_2H)^1(OH)^3$	11
	Phenol	C_6H_5OH	11
	2-4 Dinitrophenol	$C_6H_4(OH)^1(NO_2)_2{}^{2-4}$	11
	1-2-3 Pyrogallol Trimethylic Ether	$C_6H_3(OCH_3)_3{}^{1-2-3}$	11
	2 Methoxybenzylcamphor	$C_{18}H_{24}O_2$	4

TABLE VI

Binary, Monoclinic, or Clinorhombic System

Holohedral Form—L^2 C P

	Name	Symbol	Reference
Enantiomorphous Hemihedrism Class: Sphenoidal C_2 L^2 (This class is mentioned by Giebe and Scheibe as being strongly piezoelectric)	Lithium Sulfate	$Li_2SO_4 \cdot H_2O$	5
	Right-Hand Tartaric Acid	$C_4H_6O_6$	1, 5
	Potassium Tartrate	$K_2C_4H_4O_6$	5
	Ammonium Tartrate	$(NH_4)_2C_4H_4O_6$	5
	Sodium Tartrate	$Na_2C_4H_4O_6$	5
	Zinc Maleate	$ZnHOC_2H_3(CO_2)_2$	11
	Saccharose	$C_{12}H_{22}O_{11}$	1, 5
	Lactose	$C_{12}H_{22}O_{11} \cdot H_2O$	13, 11
	α Rhamnose	$C_6H_{12}O_5 \cdot H_2O$	13, 4
	Glucosamine Hydrochlorate	$C_6H_{11}NH_2O_5 \cdot HCl$	13
	Quercite	$C_6H_7(OH)_5$	11
	Isohydrobenzoin	$C_{14}H_{14}O_2$	15
	Basic Cinchonidine Sulfate	$2C_{19}H_{22}ON_2, SO_4H_2 \cdot 5H_2O$	4
	Hematoxylin Trihydrate	$C_{16}H_{14}O_6 \cdot 3H_2O$	13
	Cinchonine	$C_{19}H_{22}ON_2$	16
Hemimorphous Hemihedrism Class: Domatic C_s P	Skolesite	$Si_3O_{10}Al_2Ca \cdot 3H_2O$	-5, 16 $+11$, 13
Holohedrism* Class: Prismatic C_{2h} L^2 C P	Ammonium Iodate	NH_4IO_3	11
	Phenylurea	$CONH_2NHC_6H_5$	11
	m. Nitrophenol	$C_6H_4(OH)^1(NO_2)^3$	11
	Xylenol	$C_6H_3(OH)^1(CH_3)^{(2-5)}$	11
	2-5 Dinitrophenol	$C_6H_3(OH)^1(NO_2)_2{}^{(2-5)}$	11
	Orcine	$C_6H_3(OH)_2{}^{1-3}(CH_3)^5$	11
	Saccharine		11
No class indicated	Calcium Ethylsulfate	$Ca(SO_3OC_2H_5)_2 \cdot 2H_2O$	11
	1-3-5 Trichloraniline	$C_6H_2(NH_2)^2Cl_3{}^{(1-3-5)}$	11
	Brucine	$C_{23}H_{26}O_4N_2 \cdot 4H_2O$	17
	Basic Cinchonine Sulfate	$2C_{19}H_{22}ON_2 \cdot SO_4H_2 \cdot 2H_2O$	4

* See footnote of Table V.

TABLE VII

Assymetric, Anorthic, or Triclinic System

Holohedral Form—C

	Name	Symbol	Reference
Hemihedrism Class: Pedial C_1 No symmetry element	Potassium Bichromate Strontium Bitartrate	$K_2Cr_2O_7$ $SrH_2(C_4H_4O_6)\cdot 4H_2O$	−13, 14 +13
Holohedrism* Class: Pinacoidal C_i C	Biuret	$NH_2\text{—}CO\text{—}NH\text{—}CO\text{—}NH_2$	+14

* See footnote of Table V.

TABLE VIII

Unclassified (No System)

Name	Symbol	Reference
Sodium Lithium Molybdate	$Na_3Li(MoO_4)_2\cdot 6H_2O$	3
Epistilbite	$CaAl_2H_4(SiO_3)_6\cdot 3H_2O$	11
Potassium Sodium Aluminooxalate	$KNa_2[Al(C_2O_4)_3]\cdot 4H_2O$	11
Potassium Iridooxalate		4
2 Cobaltitriethylene Diamine		4
Sodium Malonate	$Na_2C_3H_2O_4\cdot H_2O$	11
Acid Calcium Maleate or Calcium Dimaleate	$CaC_4H_4O_5,\ H_2C_4H_4O_5,\ 5H_2O$	11
Mercury Cyanate	$Hg(CNO)_2$	14
Ethyl Cyanurate	$(CNOC_2H_5)_3$	14
Guanidonium Iodide	CH_3N_3I	19
Acetylacetic Acid Anilide	$CH_3COCH_2CONHC_6H_5$	11
Anilidoisobutyronitrile	$H_3C\text{—}C(NH\text{—}C_6H_5)\text{—}C{\equiv}N$	11
m. Nitroxylene	$C_6H_3(CH_3)_2^{1-3}(NO_2)^5$	11
p. Benzyl Nitrochloride	$C_6H_4(CH_2Cl)(NO_2)^4$	11
m. Chloracetanilide	$C_6H_4ClNHCOCH^3$	17
m. Aminophenol	$C_6H_4(OH)^1(NH_2)^3$	11
6 Nitro O Cresol	$C_6H_3(OH)^1(CH_3)^2(NO_2)^6$	11
6 Chloro 2-4 Dinitrophenol	$C_6H_2(OH)^1(NO_2)_2^{2-4}Cl^6$	11
p. Anisidine	$C_6H_4(OCH_3)^1(NH_2)^4$	11
Anisidine Formaldehyde	$C_6H_4(OCH_3)^1(N{=}CH_2)^4$	11
Methoxyphtalimide	$C_6H_3(CONHCO)^{(1-2)}(OCH_3)^4$	11
2 Ethoxynaphthalene	$C_{10}H_7OC_2H_5$	11
Anthraquinone Sodium Monosulfonate	$NaSO_3C_{14}H_7O_2(H_2O?)$	11
β Azidoacetoglucose	$C_6H_{11}O_5N_3$	11
α Oxyquinoleine	$C_9H_6N(OH)$	11
p. Aminophenol	$C_6H_4(OH)^1(NH_2)^4$	11
Orexine Hydrochlorate	$C_{14}H_{13}N_2HCl$	11
Iodoform Cyclohexane	$CHI_3C_6H_{12}$	11
pp. Tetramethyldiaminodyphenylmethane	$[(CH_3)_2N\text{—}C_6H_4]_2CH_2$	11
Quebrachite	$C_6H_6(OH)_5OCH_3$	4
Patchouli Camphor	$C_{15}H_{25}OH$	3
p. Chlorated Benzal Camphor	$C_{17}H_{19}OCl$	4
p. Dimethylaminobenzal Camphor	$C_{19}H_{25}ON$	4
Dimethylcampholamide	$C_{12}H_{23}ON$	4
Phenyl Fenchol	$C_{16}H_{22}O$	4
1 Ephedrine Hydrochlorate	$C_{10}H_{15}ON\cdot HCl$	11
Ephedrine and Right-Handed Pseudococain Right-Handed Tartrate	$C_{31}H_{42}O_7N_2\cdot H_2O$	11
Cinchoniline Hydrochlorate	$C_{19}H_{22}ON_2\cdot HCl$	4
Cinchoniline	$C_{19}H_{22}ON_2$	4
Ureoquinine Hydrochlorate	$C_{20}H_{24}O_2N_2\cdot 2HCl\cdot CH_4ON_2\cdot 5H_2O$	4
Quinine Hydrobiiodate	$C_{20}H_{24}O_2N_2\cdot 2HI\cdot 5H_2O$	4
Quinine Hydrobromate	$C_{20}H_{24}O_2N_2\cdot HBr\cdot H_2O$	4
Cinchonine Hydrate		4
Cinchonine Chlorozincate		4
Oxycinchonine α Chlorocadmiate		4
C Luteosterone	$C_{21}H_{30}O_2$	20
D Luteosterone	$C_{21}H_{30}O_2+H_2O$	20

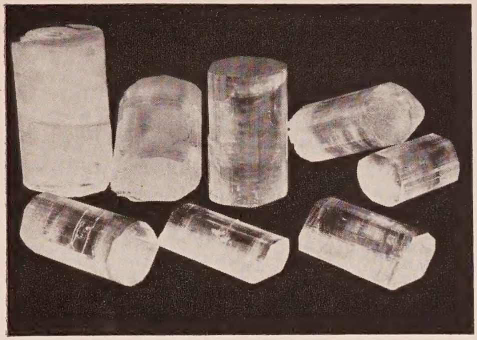


Fig. 2—Large Rochelle salt crystals.

the absence of a sign means that the substance is piezoelectric.

We mentioned negative results only when they referred to substances belonging to classes susceptible of exhibiting piezoelectricity (and only for the first papers, because of the large number of substances investigated later on). It can be seen that particularly the barium, strontium, and lead nitrates, which exhibit tetartohedrism of the cubic system, have not proved to be piezoelectric. The phenomenon may be too weakly evidenced to be detected in the samples studied, or, as Bouasse remarks, "dissymmetry of the medium does not necessarily cause a dissymmetry of phenomena." The only piezoelectric nitrate known at present is rubidium nitrate; according to Groth's treatise, it would crystallise under the bipyramidal form of the orthorhombic system and this indication should be revised since it is not consistent with the existence of piezoelectricity. Sometimes the results obtained on the same substance vary from one investigator to another, and these discrepancies are probably a consequence of the qualities of the samples examined.

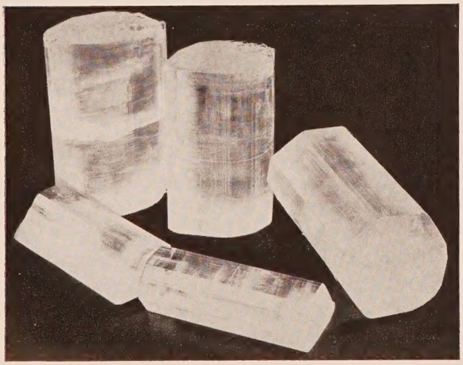


Fig. 3—Large Rochelle salt crystals.

As observed by Lucas,[4] there seems to be no exception to the fact that all substances endowed with rotatory power are piezoelectric. Hettich and Steinmetz[11] have observed, on α and β methylglucoside that the intensity of the piezoelectric phenomenon varied in the same direction as the rotatory-power value, but this observation cannot be generalized for lack of experimental data.

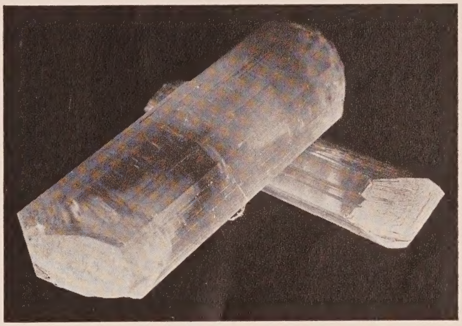


Fig. 4—Magnesium sulfate crystals.

Until now, natural crystals have been mostly used; quartz, and, to some extent, tourmaline. Among artificial crystals, Rochelle salt alone has been well investigated, and has had applications because of its remarkable piezoelectric properties.

7 *Artificial Crystals*

An examination of the above tables shows that if natural crystals happened to be lacking, a considerable number of piezoelectric substances would still be available. Their properties are still little known; in most cases lists have been drawn up without any measurements being made. This is easily explained: this kind of qualitative research requires only minute crystal fragments, whereas quantitative determinations require fine specimens the preparation of which is very often lengthy and painstaking.

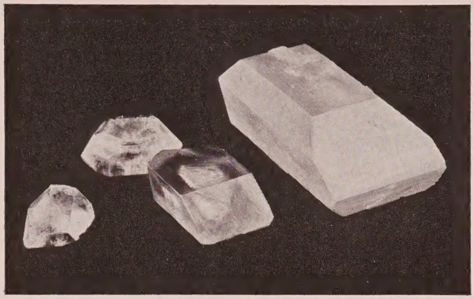


Fig. 5—A large saccharose crystal is at the right and three benzophenone crystals are at the left.

Experience shows that for the preparation of large crystals satisfactory results are obtained by evaporation or very slow cooling of a saturated solution at rest containing a few germs regularly arranged. The crystal can also be moved, slowly, in the slightly supersaturated solution, or an evaporation or a controlled cooling of a saturated solution may be effected, the solution being stirred with care and primed with a few germs. This last method may be applied to industrial preparation as fine crystals can be obtained in a relatively short time.

Figs. 2 and 3 show large Rochelle salt crystals and Fig. 4 magnesium sulfate crystals; Fig. 5, on the right, shows a large saccharose crystal and, on the left, three benzophenone crystals; Figs. 6 and 7 show sodium chlorate crystals.

7.1 ROCHELLE SALT

Rochelle salt crystallizes in the orthorhombic system. It has three orthogonal crystallographic axes, *abc* or *XYZ*. Fig. 8 represents a complete crystal; Fig. 9 a crystal incompletely developed along the *a* axis.

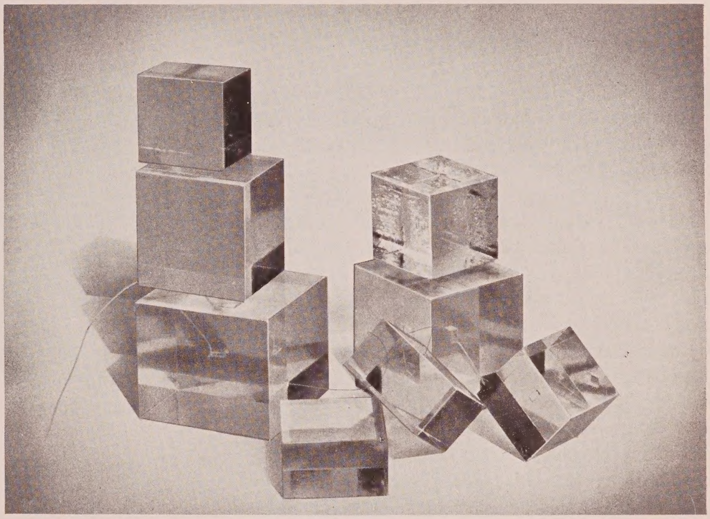


Fig. 7—Sodium chlorate crystals.

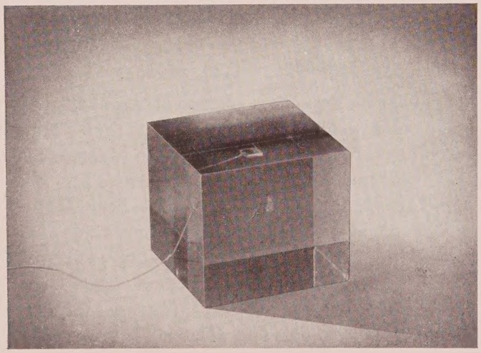


Fig. 6—Sodium chlorate crystal.

This substance has drawn the attention of scientists by the abnormal variations of its dielectric constant and of its piezoelectric modulus when the temperature varies. Mueller[25] observed that the dielectric anomalies appear only with fields whose directions are along the *a* axis. They resemble those of magnetism in ferromagnetic substances but there are two Curie points, −18 degrees and 23.7 degrees centigrade. The piezoelectric effect varies in the same manner as the dielectric polarization.

Neumann and Voigt[26] have shown that because of the symmetry of the crystal, 15 of the 18 piezoelectric moduli must be zero. Measurements have been made by Pockels,[27] Valasek,[28] Schwartz,[29] and Körner.[30] This last physicist indicated the following values for the piezoelectric moduli at 21 degrees centigrade:

$d_{14}=3\cdot 10^{-5}$ electrostatic cgs unit/dyne,
$d_{23}=0.39\cdot 10^{-5}$,
$d_{36}=0.33\cdot 10^{-5}$.

Schulwass Ssorokina[31] measured the variations of d_{14} with temperature. See Table IX. This d_{14} modulus is extraordinarily large when compared to the moduli of other piezoelectric substances. In particular, the d_{11} modulus of quartz is $6.5\cdot 10^{-8}$. This property gives Rochelle salt a very marked superiority over quartz for certain uses, although its mechanical properties are not as remarkable as those of the natural crystal.

TABLE IX

Temperature, Degrees Centigrade	$d_{14}\times 10^5$	Temperature, Degrees Centigrade	$d_{14}\times 10^5$
−30	0.6	+ 5	26.4
−20	1.6	+10	26.4
−15	12.0	+20	26.4
−10	18.4	+25	2.0
− 5	22.8	+30	1.0
0	25.0		

Certain physicists have measured the piezoelectric moduli during the vibration of the crystal. They found values which differ appreciably from those obtained by static methods. This discrepancy may be attributed to the high relaxation time of these crystals which does not allow them to reach their static conditions when an alternating voltage is applied. The experiments of Mikhailov[32] were made on a Rochelle salt crystal oriented so as to form an angle of 45 degrees with the b and c axes (Fig. 10). An electric field E_x is applied along the a axis; the value of modulus d_{14} as determined during the crystal oscillation is $4.5\cdot 10^{-6}$ absolute units at 18.5 degrees centigrade. Mason[33] uses fields below 40 volts per centimeter and frequencies of the order of 80 kilocycles (as shown by Norgorden,[34] the results obtained are independent of the frequency provided the frequency is higher than 100 cycles). The maximum of d_{14}, which occurs about 24 degrees centigrade, is $1600\cdot 10^{-8}$ centimeter-gram-second units. The work of Schulwass Ssorokina[35] on this subject should also be mentioned.

According to Stamford,[36] piezoelectric resonators exhibiting only a pure longitudinal vibration are obtained by cutting the samples in such a

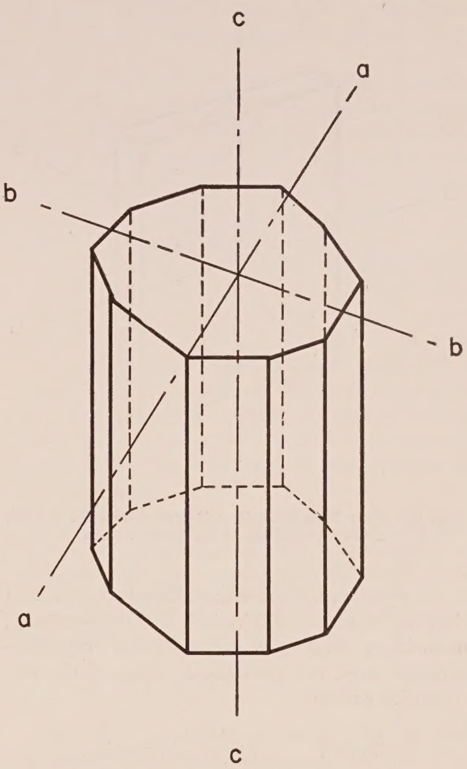


Fig. 8—A complete Rochelle salt crystal having three orthogonal crystallographic axes *abc*.

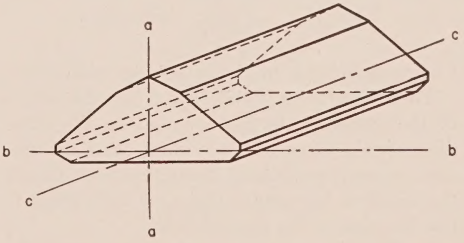


Fig. 9—Rochelle salt crystal incompletely developed along the *a* axis.

way that one axis coincides with the crystallographic axis and the two others are at 45 degrees to the other crystal axes, the field being applied parallel to the axis which coincides.

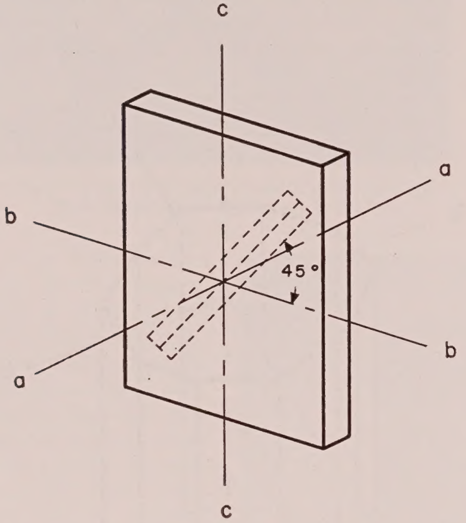


Fig. 10—Plate from Rochelle salt crystal oriented to form a 45-degree angle to the *b* and *c* axes.

If three samples are considered (I, II, III) having their axes OX', OY', OZ', respectively, coincident with the corresponding crystallographic axes, the theoretical values of the frequencies will be

$$\text{Type I} \qquad f=\frac{1900}{l} \text{ kilocycles,}$$

$$\text{Type II} \qquad f=\frac{1286}{l} \text{ kilocycles,}$$

$$\text{Type III} \qquad f=\frac{1975}{l} \text{ kilocycles,}$$

l being the length of the sample in millimeters.

These results are valid only when the length of the sample is large compared to the other dimensions; otherwise a second resonant frequency would result from standing waves through the width of the sample and at a right angle to the direction of the applied field.

As there is no action in the direction of the applied field, the contacts to the electrodes do not increase the damping of the crystal as much as in the case of quartz; similarly, the use of an air gap between the crystal and the electrode does not increase the damping, as observed by Debye in the case of quartz.

The above-mentioned type of cut is the most widely used. Cady, however, suggested the use of X-, Y-, and Z-cut Rochelle salt elements with their length parallel to one of the orthogonal axes of the substance. The firm, Le Matériel Téléphonique, recommends in a patent[37] the use of X-, Y-, and Z-cut crystals whose principal plane faces are parallel to one axis, the planes of these faces being at 45 degrees to the two other axes (Fig. 11).

Such elements can work in a high-frequency shear vibration along their smallest dimension and their piezoelectric and dielectric constants are relatively stable when the temperature varies. The patent shows how to calculate the resonance frequencies in terms of the dimensions and of the elastic constants of the elements.

Golay[38] built an electrometer based on the piezoelectric properties of Rochelle salt; he observed deflections proportional to the voltage and independent of frequency, provided the frequency range is far from the frequencies of mechanical resonance.

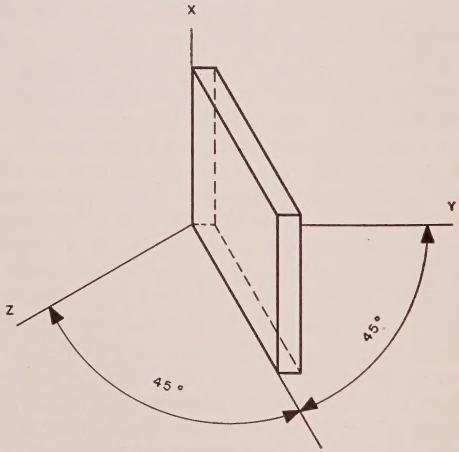


Fig. 11—Rochelle salt plate cut along the *X* axis with its principal plane faces at 45 degrees to the *Y* and *Z* axes. Similar cuts may have the principal face along the *Y* or *Z* axis and at 45 degrees to the other two axes.

Rochelle salt is mostly used for microphones, phonographs, pickup devices, loudspeakers, relays, and light valves. It may also be useful in filter construction[39] where its substitution for quartz, offers several advantages:

> Owing to the properties of quartz, the filter bandwidth is predetermined unless special devices are used, such as the addition of inductances. Filters consisting only of oscillating crystals allow a bandwidth of 0.8 percent of the mean filter frequency, as a maximum. For a 300-kilocycle frequency, the maximum bandwidth, therefore, is 2400 cycles. The result is that for the transmission of a speech band, quartz filters can be used only to about 300 kilocycles and the limit of carrier telephony, 5 kilocycles approximately, is not accessible to such filters.
>
> The advantage of Rochelle salt filters is that while having steep sides like quartz filters, they have an appreciably wider band, reaching a maximum of about 6 percent of the mean frequency. Therefore the 2400-cycle bandwidth can still be obtained down to a mean frequency of 40 kilocycles only. This property may be attributed to the fact that Rochelle salt has a higher piezoelectric frequency than quartz and, at the same time, a lower modulus of elasticity.
>
> The dielectric constant of Rochelle salt is higher than that of quartz. This property is somewhat offset, it is true, by the widening of the band, but it offers the advantage that the additional capacitors which may be necessary have an order of magnitude more readily realizable. Another advantage is the small space expansion of the oscillating crystals which, in turn, must be attributed first of all to the lower modulus of elasticity.

Rochelle salt crystallizes with four water molecules; Holden, Mason, and their collaborators[40, 41] suggested replacing this water by heavy water (deuterium oxide). They observed that the higher Curie point changes from 23.5 degrees to 34.5 degrees centigrade and the lower Curie point from -19 to -24 degrees centigrade. The elastic, dielectric, and piezoelectric properties are practically unchanged.

The firm, Electrical Research Products,[42] derived a practical application from these observations. They use

> Rochelle salt plates (whose maximum efficiency is between 16 and 28 degrees centigrade) between which plates of the same salt crystallized from deuterium oxide are inserted (maximum efficiency between 28 and 42 degrees centigrade). The composite device thus obtained is made sensitive over the entire range from 16 to 42 degrees centigrade, which is much wider than that of the Rochelle salt devices of the known type.

The properties of sodium-rubidium tartarate have also been used. Electrical Research Products suggested inserting plates of this salt in the above-described composite systems. Since the Curie point of this salt is lower than that of Rochelle salt (14 degrees centigrade), the temperature range of the device is widened further. The variations with temperature of the piezoelectric modulus are much lower in the case of the composite system than in the case of a single crystal.

The firm, Le Matériel Téléphonique, mentions in one of its patents,[43] that the frequency change with temperature or the frequency-temperature coefficient goes through a transition point at the Curie-point temperature (about 23.6 degrees centigrade). Below this temperature it is approximately -1200; above it, it is about -500. They suggest the introduction of 0.3 percent isomorphous sodium and rubidium tartarate which decreases the Curie point of the resultant crystal to 7 degrees centigrade, i.e., below room temperatures. The frequency-temperature coefficient is relatively constant above the new Curie point and its value is about one-half of that of Rochelle salt. The same patent mentions that the negative temperature coefficient can be compensated for by the construction

> of a resonance or vibrating device consisting of a piezoelectric part and a nonpiezoelectric part mechanically associated to vibrate as a single element; the nonpiezoelectric part has a frequency-temperature coefficient of an opposite sign and its magnitude is such that the combined element has a coefficient which is practically zero.

7.2 OTHER ARTIFICIAL CRYSTALS

We have very little information on the other artificial piezoelectric substances.

Busch[44] observed that KH_2PO_4 and KH_2AsO_4 offer the same anomalies as Rochelle salt. Ludy[45] indicated that the value of d_{36} increases slowly from $50 \cdot 10^{-8}$ at room temperature, to $1000 \cdot 10^{-8}$ at 143 degrees centigrade (absolute), then increases very rapidly around the lower Curie point, to reach a maximum of $32\,500 \cdot 10^{-8}$ at 123 degrees centigrade (absolute).

The piezoelectric constant of NH_4Cl was determined by Bahr and Engl.[21] The piezoelectric effect starts at -30.4 degrees centigrade and reaches its maximum value at -32.5 degrees centigrade. As the substance crystallizes in the cubic system, a single modulus is found d_{14} whose value at -32.5 degrees centigrade is $0.337 \cdot 10^{-8}$;

it is twenty times smaller than that of the d_{11} modulus for quartz.

Sodium chlorate, which similarly crystallizes in the cubic system, has also a single modulus d_{14}. It was measured by Pockels[27] who gives its value as $-4.84 \cdot 10^{-8}$.

Van der Veen,[3] then Knoll,[46] determined the d_{14} modulus for blende (which also belongs to the cubic system). This last physicist gives a value $d_{14} = (9.7 \pm 0.15) \times 10^{-8}$ centimeter-gram-second units.

Pawlik[47] showed that a properly cut saccharose plate can be excited to oscillations. Table X gives the values of the moduli for right-hand tartaric acid and saccharose, which crystallize in the same system. For comparison purposes, the moduli for quartz and tourmaline follow the table.

TABLE X

Moduli	Right-Hand Tartaric Acid (Tamaru[48])	Saccharose (Holman[49])
d_{14}	-24	$+1.2$
d_{15}	$+28$	-1.2
d_{24}	$+28.5$	-7.2
d_{25}	-36.5	-3.7
d_{31}	$+1.9$	$+2.2$
d_{32}	$+5.9$	$+4.4$
d_{33}	$+6.4$	$-10.$
d_{36}	$+3.8$	-2.6

Quartz: $d_{11} = 6.36$,[26] $= 6.94$,[51, 52] $= 6.9$ (Voigt and Freedericksz); and $d_{14} = 1.69$.[26]
Tourmaline: $d_{15} = +11$;[53] $d_{22} = +0.68$;[53] $d_{31} = +0.74$;[53] $d_{33} = -5.41$;[54] and $+5.78$.[53]

All moduli are given in 10^{-8} centimeter-gram-second electrostatic units.

Finally, in his 1911 paper, Van der Veen[3] gives the value of charge E_x in electrostatic units, caused by a 1-kilogram load for the substances listed in Table XI.

TABLE XI

Substance	Symbol	Charge E_x in *C-G-S* Electrostatic Units
Benzil	$C_6H_5COCOC_6H_5$	0.237
Sodium-Lithium Molybdate	$Na_3Li(MoO_4)_2.6H_2O$	about 0.140
Barium Antimonyl Tartarate	$Ba(SbO)_2(C_4H_4O_6)_2.H_2O$	0.109
Rubidium Tartarate	$Rb_2C_4H_4O_6$	about 0.080
Quartz		0.062
Tourmaline		0.053
Blende		0.047
Patchouli Camphor	$C_{15}H_{25}OH$	0.0014
Diamond		0.000

We cannot end this study, dedicated to piezoelectric substances obtained artificially, without mentioning research work done in quite a different manner and outlined in two patents obtained in 1929. The authors of these patents tried to prepare substances exhibiting piezoelectric properties (and, in some cases, rotatory power), by orienting an infinity of small hemihedric crystals.

Telefunken[55] uses slabs cut from quartz crystals or amorphous quartz and suggests loosening the links between molecules by very intense heating. The slabs are then exposed to the action of a very strong electric field, the intensity of which is increased gradually during cooling. This would give piezoelectric slabs.

Freedman and Rickets[56] suggest orienting small crystals in a properly chosen dielectric medium (preferably with a high dielectric constant), such as waxes, greases or hydrogenated oils, natural resins, celluloid; one or several optically active substances are included, like zinc oxide, neutral cinchonin or cuprein sulfate, dicamphohexanedione, or peroxide of dicamphohexanedione. In some cases these substances may be partially or completely replaced by cholesterol or phytosterol, which are present in oils subjected to low-temperature hydrogenation, of by pinnaric acid or β sylvic acid existing in resins or colophony.

Orientation is obtained by electrical or mechanical methods. According to the authors, large disks could be prepared in this manner, for microphones and light interrupters used in television.

These patents show that as early as 1929, engineers were already trying to replace natural crystals by artificial products. Their technique does not seem to have been kept up, but since that time the question has been studied unceasingly. Rochelle salt, which is well known because of the many theoretical investigations of it, appeared as a likely substitute for quartz in most of its uses. The efforts of technicians were therefore directed towards adapting this substance to industrial needs.

There can be no question of stopping here; the vast field of artificial piezoelectric bodies has been explored to only a limited extent; the study of the properties of these substances is of great theoretical and practical importance and it is to be hoped that investigators, physicists, and chemists will develop an interest in them, their collaboration being indispensable in bringing this work to a satisfactory conclusion.

8 Bibliography

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Electron Trajectories in a Plane Single-Anode Magnetron—A General Result

By L. BRILLOUIN

Harvard University, Cambridge, Massachusetts
Consulting Engineer, Federal Telecommunication Laboratories, Inc., New York, New York

A GENERAL theorem for electron trajectories in a plane diode (with space charge) was proved in a previous paper.[1] It stated that electron trajectories could never cross each other unless some time intervals with negative current might cause back bombardment of the cathode.

A similar theorem is proved in the present paper for the case of a plane magnetron. There is however a new condition to be added; that saturation should never be reached. Some examples are given, together with a complete solution for electronic motion, when the current through the magnetron is any given function of time.

• • •

1 General Theorem for Electron Trajectories

Electron trajectories in a plane diode with space charge were investigated by the author in two recent papers[1, 2] and a general theorem was stated:

> In a plane diode with hot cathode and space charge, operated under arbitrary voltage variation as a function of time, electron trajectories never cross each other, provided the current never becomes negative.

The present paper is devoted to a similar discussion for a plane single-anode magnetron, having exactly the same structure as the plane diode but with a strong magnetic field perpendicular to the plane of the figure.

Using a method very similar to the one previously followed,[1] the *same theorem will be proved for magnetrons*, but with the *additional condition* that *saturation should never be reached.* In other words, if we apply an arbitrary voltage variation on the anode of a plane magnetron, electron trajectories will never cross each other, provided:

I. Current never becomes negative.

II. Current remains space-charge limited and saturation current is never obtained.

This second condition represents the only difference between the diode and magnetron problems. In the case of a plane diode, we found condition I, but II did not occur and it was immaterial whether the current was space-charge limited or temperature limited.

The importance of the preceding theorem will be better understood if one remembers that, in connection with the theory of magnetrons, two different types of solution have been proposed by different authors.

A. *Double-Stream Motions*, with some electrons flowing away from the cathode while others are striking back upon the cathode.

B. *Single-Stream Motion*, where electrons are flying only in one direction, from cathode to anode.

In the limiting case of zero anode current, the solution B yields electron trajectories parallel to cathode and anode, with no emission from the cathode. This type of solution was proposed some years ago by the author[3–5] but the discussion of electron trajectories was not rigorous and contained some approximations that were not very satisfactory. The theory developed in the present paper is rigorous and a special discussion is given of electron trajectories under cutoff conditions, leading to the B single-stream motions.

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[1] L. Brillouin, "Electron Trajectories in a Plane Diode—A General Result," *Electrical Communication*, v. 22, n. 3, pp. 212–216; 1945.

[2] L. Brillouin, "Transit Time and Space-Charge in a Plane Diode," *Electrical Communication*, v. 22, n. 2, pp. 110–123; 1944.

[3] L. Brillouin, "Theory of the Magnetron," *Physical Review*, Part I, v. 60, pp. 385–396; September, 1941: Part II, v. 62, pp. 166–167; August, 1943: and Part III, v. 63, pp. 127–136; February, 1943.

[4] L. Brillouin, "Theory of the Magnetron," *Electrical Communication*, v. 20, n. 2, pp. 112–121; 1941.

[5] L. Brillouin, "Practical Results from Theoretical Studies of Magnetrons," *Proceedings of the I.R.E.*, v. 32, pp. 216–230; April, 1944.

The theory of electron trajectories in a plane magnetron under steady conditions has been known for some years.[6] Its extension to variable conditions is contained in the present papers and can be applied to a number of problems of single-anode plane magnetrons. Similar methods could also be extended to cylindrical single-anode magnetrons, but for some serious mathematical difficulties.

The theory of multianode magnetrons has not been worked out to the same extent and still relies on rather delicate approximations.

2 Electron Trajectories in the Magnetron

The magnetron has a plane cathode $(y=0)$ and a plane anode $(y=d)$ with a magnetic field H in the z direction. Electrons move in the xy plane. The x motion is readily integrable.

$$\left.\begin{aligned} \ddot{x} &= -2\omega_H \dot{y}, \\ \omega_H &= -\tfrac{1}{2}\mu_0 \frac{e}{m} H, \\ \dot{x} &= -2\omega_H y, \\ \omega_H&, \text{ Larmor frequency } e<0. \end{aligned}\right\} \quad (1)$$

There is no integration constant since the velocity $\dot{x}$ must be zero on the cathode $y=0$.

The y motion is given by

$$\ddot{y} = 2\omega_H \dot{x} + \frac{e}{m}E = -4\omega_H^2 y + \frac{e}{m}E, \qquad (2)$$

where E is the electric field acting on the electron at time t. We assume all quantities to depend on y and t but not on x.

Here we use Llewellyn's method of integration. We first write the total current

$$-I = \rho \dot{y} + \epsilon_0 \frac{\partial E}{\partial t}, \qquad (3)$$

where $\rho < 0$ is the electric density. The minus sign yields a positive current when electrons move toward the anode and $\dot{y} > 0$.

We assume a B motion (single stream) and look for conditions that will maintain this type of solution. The Poisson relation

$$\epsilon_0 \frac{\partial E}{\partial y} = \rho \qquad (4)$$

together with (3) yields

$$\frac{dE}{dt} = \frac{\partial E}{\partial t} + \dot{y}\frac{\partial E}{\partial y} = -\frac{I}{\epsilon_0}, \qquad (5)$$

where d/dt is taken along the path of an electron. The important point is that the *total current* I is a *function of* t *only* and *does not depend on* y. This results directly from Maxwell's theory of the displacement current.

Let us see as parameters the *time* t_0 at which an *electron* leaves the cathode, and the *transit time* τ. The time t at which this electron is observed is

$$t = t_0 + \tau. \qquad (6)$$

On the cathode we have an electric field $E_0(t_0)$

$$\left.\begin{aligned} &E_0 < 0, \\ &eE_0 > 0, \text{ saturation current}, \\ &E_0 = 0, \text{ space-charge-limited current}. \end{aligned}\right\} \quad (7)$$

Integrating (5) we obtain the electric field acting on the electron at time t,

$$E(t_0, t) = E_0(t_0) + F(t) - F(t_0),$$

$$F(t) = -\frac{1}{\epsilon_0}\int_0^t I(t)dt. \qquad (8)$$

This gives the field in the whole region occupied by the space charge, as we always follow the motion of an electron.

On the cathode, we have the usual initial conditions of emission without velocity

$$y=0, \quad \dot{y}=0 \quad \text{at} \quad t=t_0, \quad \tau=0. \qquad (9)$$

Let us now assume that we know a special solution $f(t)$ of

$$\ddot{f} + 4\omega_H^2 f = \frac{e}{m}F(t). \qquad (10)$$

The general solution of (2) and (8) can be immediately written

$$y(t_0, t) = f(t) - \frac{e}{m4\omega_H^2}[F(t_0) - E_0(t_0)] + C\cos 2\omega_H\tau + S\sin 2\omega_H\tau, \qquad (11)$$

where the C and S coefficients must be adjusted to satisfy the initial conditions of (9).

$$\left.\begin{aligned} C &= -f(t_0) + \frac{e}{m4\omega_H^2}[F(t_0) - E_0(t_0)], \\ S &= -\frac{1}{2\omega_H}\dot{f}(t_0). \end{aligned}\right\} \quad (12)$$

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[6] E. B. Moullin, *Proceedings Cambridge Philosophical Society*, v. 36, p. 94; 1940.

The very important point in this solution is that it includes terms of frequency $2\omega_H$ and terms in f or F that will contain the applied frequency ω in case of periodic operation. Hence beats between both frequencies will play an important role in shaping the electronic trajectories.

To investigate the relative position of the trajectories, we compute

$$\left[\frac{\partial y}{\partial t_0}\right]_{t\ \text{const.}} = -\frac{e}{m4\omega_H^2}(\dot{F}_0-\dot{E}_0) + \left(\frac{\partial C}{\partial t_0}-2\omega_H S\right)\cos 2\omega_H\tau + \left(\frac{\partial S}{\partial t_0}+2\omega_H C\right)\sin 2\omega_H\tau. \tag{13}$$

If this expression stays negative, the successive electronic layers remain stratified and never intercross each other.

According to (12)

$$\left.\begin{aligned} \frac{\partial S}{\partial t_0}+2\omega_H C &= -\frac{e}{m2\omega_H}E_0,\\ \frac{\partial C}{\partial t_0}-2\omega_H S &= \frac{e}{m4\omega_H^2}(\dot{F}_0-\dot{E}_0),\\ &= -\frac{e}{m\epsilon_0 4\omega_H^2}i(t_0), \end{aligned}\right\} \tag{14}$$

since

$$\begin{aligned} \dot{F}_0-\dot{E}_0 &= -\frac{1}{\epsilon_0}[I(t_0)+\epsilon_0\dot{E}_0],\\ &= -\frac{i(t_0)}{\epsilon_0}, \end{aligned}$$

where $i(t_0)$ is the electronic part of the current to the cathode at t_0 while $\epsilon_0\dot{E}_0$ is the displacement current to the cathode. Finally

$$\left(\frac{\partial y}{\partial t_0}\right)_{t\ \text{const.}} = \frac{e}{m\epsilon_0 4\omega_H^2}i(t_0)[1-\cos 2\omega_H\tau] - \frac{e}{m2\omega_H}E_0\sin 2\omega_H\tau. \tag{15}$$

This is the formula on which we base our discussion. Equations (11) and (12) yield the general solution for electron motion in a plane magnetron operated under arbitrary conditions. This solution can be used as long as the motion is of the single-stream B type. The static solution of Moullin[6] corresponds to the case

$$\left.\begin{aligned} &I=I_0\ \text{constant}, && f=-\frac{eI_0}{m\epsilon_0 4\omega_H^2}t,\\ &F=-\frac{I_0}{\epsilon_0}t, && E_0=0. \end{aligned}\right\} \tag{16}$$

3 "*B Single-Stream*" versus "*A Double-Stream*" Motions

Let us now discuss the conditions under which

$$\left(\frac{\partial y}{\partial t_0}\right)_{t\ \text{const.}} < 0 \tag{17}$$

and "B Single-Stream" motions are safeguarded. The electronic part $i(t_0)$ of the current at the moment t_0 when our electron leaves the cathode is obviously a positive quantity. Since e is negative for electrons, the first term in (15) is always negative, but the second term is oscillating. Hence the result predicted in section 1 above.

For space-charge-limited current ($E_0=0$), the derivative (17) remains negative, and electron trajectories do not cross each other. A single-stream B motion, once started, maintains its character unless some time intervals with negative currents may occur, in which case the same difficulties encountered for the diode will be found again for the magnetron.[1] New electrons may leave the cathode at a moment when former electrons are still falling back upon it.

Assuming no negative currents, we want to investigate the case of *saturation* current.

$$\left.\begin{aligned} E_0 &< 0,\\ eE_0 &> 0. \end{aligned}\right\} \tag{18}$$

The derivative (15) is still negative at the beginning (τ small) but it will become positive when

$$2\omega_H\tau = 2\pi-\epsilon, \tag{19}$$

as the first term is small, while the second one is positive. Intercrossing of trajectories will occur near the end of the first "double Larmor" period and the motion will lose the single-stream type and become "double stream."

This proves the general theorem stated in the first section of this paper.

Assuming again a space-charge-limited current ($E_0=0$), we note that the derivative (15) is zero when

$$2\omega_H\tau = 2\pi,\ 4\pi,\ \cdots,\ 2n\pi, \tag{20}$$

which means that electron trajectories touch each other without crossing and an *infinite space-charge layer* is obtained. This is a peculiar feature of the magnetron, and nothing similar was found in the usual diode.

4 Examples of Electron Trajectories in Magnetrons

The method described in section 2 above will be applied to the special example of a rectangular current impulse

$$I=\begin{cases} 0 & \text{if} \quad t<0, \\ I_0 & \quad 0<t<T, \\ 0 & \quad T<t. \end{cases} \tag{21}$$

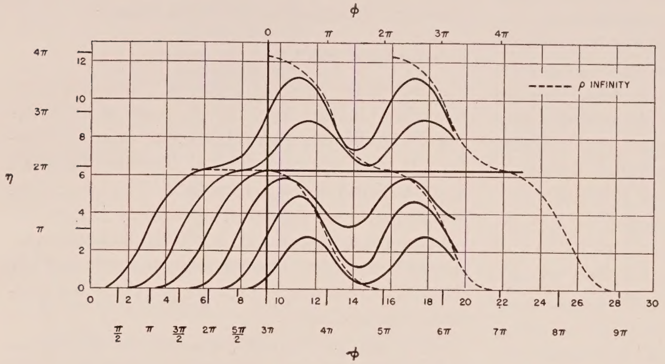

Fig. 1

We start with a function $F(t)$ as defined in (8)

$$F(t)=\begin{cases} 0 & \text{if} \quad t<0, \\ -\dfrac{I_0}{\epsilon}t & \quad 0<t<T, \\ -\dfrac{I_0}{\epsilon}T & \quad T<t. \end{cases} \tag{22}$$

Next we build the $f(t)$ function of (10)

$$(t)=\begin{cases} 0 & \text{if } t<0, \\ Dt & 0<t<T, \\ D\left[T+\dfrac{1}{2\omega_H}\sin 2\omega_H(t-T)\right] & T<t, \end{cases} \tag{23}$$

$$D=-\frac{e}{m\epsilon_0}\cdot\frac{I_0}{4\omega_H^2}.$$

The sine term in the last formula is needed to insure continuity of f and its time derivative at T. Finally we obtain the trajectories

$$\eta=\begin{cases} \varphi-\varphi_0-\sin(\varphi-\varphi_0) & \text{if } 0<t_0<t<T, \\ \Phi-\varphi_0-\sin(\varphi-\varphi_0)-\sin(\varphi-\Phi) & \\ & \text{if } T<t, \end{cases} \tag{24}$$

where $\eta=2\omega_H\dfrac{y}{D}$, $\varphi_0=2\omega_H t_0$,

$\varphi=2\omega_H t$, $\Phi=2\omega_H T$, and

t_0 time of emission of an electron $0<t_0<T$. These trajectories have been drawn in Fig. 1, which corresponds to the case

$$\Phi=2\omega_H T=3\pi$$

and shows large oscillations of the space-charge border. With $\Phi=2\pi, 4\pi, \cdots$, the border of the space charge would remain in a fixed position, but large undamped oscillations would still be observed along the trajectories of electrons inside the space charge. In all cases, the anode voltage when computed is found to exhibit oscillations, despite the fact that after T the total current remains identically naught (21). This means that our example corresponds to experimental conditions with an infinite external resistance. In an actual experiment, the anode would be connected to the battery through a large but finite resistance. This would not change the beginning of the discharge, but the oscillations in voltage would draw a small oscillating current from the magnetron, and energy dissipation in the external resistance would progressively damp out the electronic space-charge oscillations. The final result would exhibit no space-charge oscillations, the electrons running parallel to cathode and anode. This is the typical B motion under no current, as described in the first section.

On the curves of Fig. 1, one recognizes the lines for infinite space charge (dashed lines), according to condition (20),

$$\varphi-\varphi_0=2\omega_H\tau=2n\pi.$$

The present paper is an abstract of a more comprehensive discussion[7] of the theory of magnetrons conducted by the author as a research project of the Applied Mathematics Group of Columbia University (Applied Mathematics Panel, National Defense Research Committee).

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[7] L. Brillouin, "Influence of Space Charge on the Bunching of Electron Beams," *Physical Review*, v. 70, pp. 187–196; September, 1942.

Selenium Rectifiers for Broadcast Radio Receivers

By EDWARD W. CHADWICK

Federal Telephone and Radio Corporation, Newark, New Jersey

FOR a given load current, and within other broad limits, the output voltage of a rectifier-filter type power supply increases with the size of the filter input capacitance. The safe forward peak-current rating of the rectifier is the major limitation of such capacitance.

The performance of many broadcast radio receivers can be improved by substituting a selenium rectifier for the commonly used vacuum diodes. The greater forward peak-current rating of the selenium rectifier, compared to the diodes, permits larger filter input capacitance, with consequent higher output voltage and reduced ripple. In addition, no filament heating power is required and operation of the selenium rectifier is instantaneous with turning on the receiver. Installation of the rectifier is simplified.

• • •

Selenium rectifiers[1–4] were first introduced in the U.S.A. by the I.T.&T. System, in 1938. They were developed further during the war years, and applications to peacetime uses are once more in mind. In many broadcast receiver designs, the input alternating voltage is limited and the selenium rectifier offers a means of increasing the operating voltages applied to the tubes.

1 *Selenium Rectifier Assembly*

A selenium rectifier plate is constructed by processing a thin layer of selenium on a nickel-plated aluminum base plate. The selenium surface is then sprayed with a low-melting-point alloy to form a front electrode. The plates are stacked over an insulated stud, screw, or eyelet, passing through a central mounting hole to make up an assembly. The plates are separated by metallic washers, and lugs are provided for electrical connections.

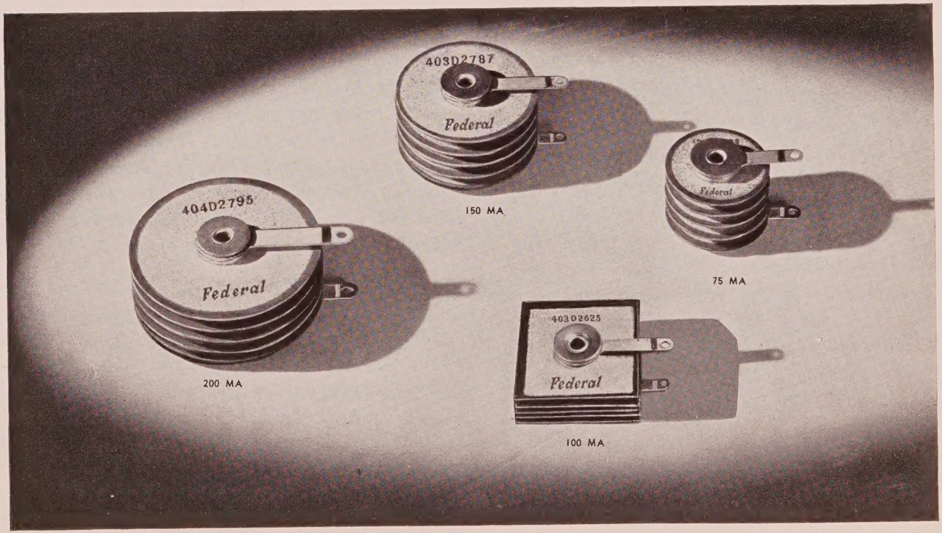

Fig. 1—Selenium rectifiers designed for use in radio broadcast receiving sets. Four sizes are illustrated, and cover a range of direct-current output from 75 to 200 milliamperes.

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[1] Carole A. Clarke, "Selenium Rectifier Characteristics, Applications and Design Factors," *Electrical Communication*, v. 20, n. 1, pp. 47–66; 1941.

[2] E. A. Richards, "The Characteristics and Applications of Selenium Rectifiers," *Journal of The Institution of Electrical Engineers*, v. 88, Part III, n. 4; December, 1941.

[3] E. Yarmack, "Selenium Rectifiers and Their Design," *Electrical Communication*, v. 20, n. 4, pp. 275–286; 1942.

[4] W. F. Bonner, "Cathodic Protection and Applications of Selenium Rectifiers," *Electrical Communication*, v. 22, n. 2, pp. 130–137; 1944. Errata, v. 22, n. 4, p. 338; 1945.

The application of an alternating voltage to the rectifier produces a much larger current flow from the base plate to the front electrode than in the opposite direction.

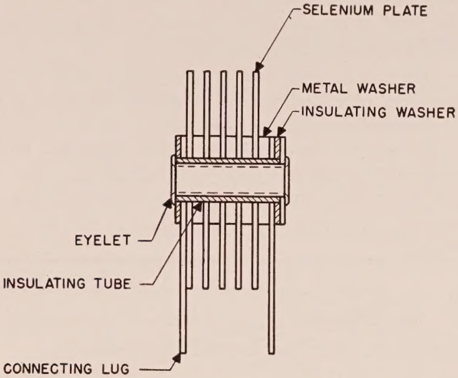


Fig. 2—Cross-sectional drawing of selenium rectifier.

2 *Electrical Design*

For satisfactory electrical design, it is necessary to know the nature of the load connected to the rectifier (resistive, capacitive, etc.), the direct current and voltage required, the ambient operating temperature, and the duty cycle. From this information, and knowledge of the electrical circuit in which the rectifier is to operate (single-phase half-wave, single-phase bridge, etc.), it is possible to determine the number and size of plates necessary, and the alternating-voltage input required. Details of the methods of rectifier design can be found in a previous technical paper.[2]

Realizing the advantages of their light-weight compact construction and effective performance, the 403D2625 rectifier, as shown in Fig. 1, was developed for use primarily in AC/DC radio receivers. For other types of receivers and electronic equipments, three additional rectifier sizes are offered. Plate voltage for these designs is obtained directly from the 117-volt alternating-current mains through a half-wave rectifier and a capacitance-input filter. The operating characteristics of these selenium rectifiers in a half-wave circuit are given in Table I.

The design of the 403D2625 (a cross-section of which is shown in Fig. 2) is an assembly of five plates, each approximately 1.25 inches square, mounted on an insulated brass eyelet, and separated from one another by metal washers 0.059 inch thick. Lugs are provided at each end of the assembly for electrical connections. The entire assembly is mounted by passing a number 6 machine or drive screw through the hollow central eyelet into the receiver chassis.

A test circuit is shown in Fig. 3 where a simple capacitance-input filter is used. An analysis of this circuit shows that the rectifier must deliver the 100-milliampere output plus the capacitance-charging current, and yet withstand a maximum peak inverse voltage of 380, which results from the addition of the line voltage of 117 volts (±10 percent) and the voltage of the charged capacitor. The voltage and current conditions in this circuit are shown in Fig. 4.

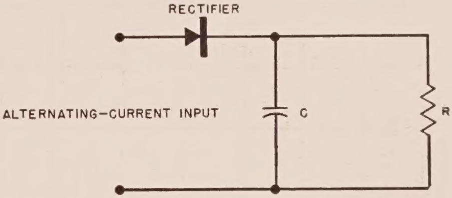


Fig. 3—Electrical test circuit for rectifier.

The performance of a selenium rectifier is unlike that of a tube, in that the operating temperature of the rectifier plate is the limiting factor in

TABLE I

Operating Characteristics as Half-Wave Rectifiers

Characteristic	402D2788	403D2625	403D2787	404D2795
Maximum R-M-S Volts	130	130	130	130
Maximum Inverse Peak Volts	380	380	380	380
Maximum R-M-S Milliamperes	220	325	425	550
Maximum Peak Milliamperes	900	1200	1200	2000
Maximum Direct-Current Output Milliamperes	75	100	150	200
Approximate Rectifier Voltage Drop, Volts	5	5	5	5
Shape of Plates	Circular	Rectangular	Circular	Circular
Diameter of Plates, Inches	1	$1\frac{1}{4} \times 1\frac{5}{32}$	$1\frac{5}{32}$	$1\frac{3}{4}$
Stack Thickness, Inches	$\frac{15}{16}$	$\frac{11}{16}$	$\frac{15}{16}$	$\frac{15}{16}$

its use; it was found necessary to limit the maximum plate-temperature rise to 40 degrees centigrade. The plate-temperature rise depends on the sum of the forward and reverse losses. It is necessary to limit these losses to two watts per plate.

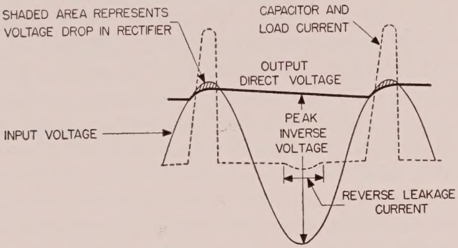


Fig. 4—Voltage and current waveforms obtained with selenium rectifier and capacitive-input filter.

Since it is the root-mean-square values of currents and voltages that determine the losses of the rectifier, the maximum peak values of forward current and inverse voltage are considerably higher, as noted in the rectifier rating.

With the aid of a calibrated cathode-ray oscilloscope, the current-voltage relationships, as indicated in Fig. 4, were determined. Due to the asymmetrical shape of the voltage and current waves involved, a graphical analysis was made to determine the root-mean-square values of the forward current and the reverse voltage. With these values established, the forward voltage drop and reverse current were determined on the basis of an equal distribution of the forward and reverse losses in the rectifier.

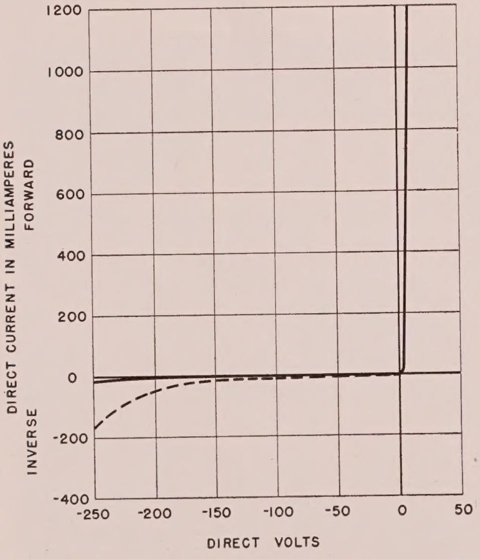


Fig. 5—Typical static curve of the forward and inverse voltage-current characteristics of the 403D2625 selenium rectifier. Dotted line represents current ×100.

Table II gives the results for a 403D2625 rectifier obtained with a nominal input of 117 volts to the circuit of Fig. 3.

A typical curve of the static forward- and reverse-current characteristics of the rectifier is shown in Fig. 5. The regulation characteristics for the circuit of Fig. 3 are shown in Fig. 6.

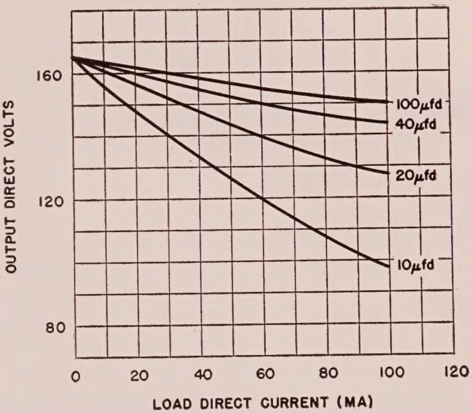


Fig. 6—Typical regulation curves using various values of filter-input capacitance with a half-wave selenium rectifier.

3 *Circuit Applications*

Figs. 7A, 7B, and 7C give typical vibrator power supplies using the selenium rectifier in half-wave, center-tapped, and bridge circuits. Fig. 8 shows a power-supply circuit for AC/DC

TABLE II

Characteristics of 403D2625 in Test Circuit

R-M-S Input Volts	117
Filter Capacitance (μfd)	80
Output Direct Current (milliamperes)	100
Inverse R-M-S Volts Across Rectifier	214
Inverse R-M-S Current Through Rectifier (milliamperes)	14
Forward R-M-S Volts Drop Across Rectifier	6.3
Forward R-M-S Current Through Rectifier (milliamperes)	508
Inverse Loss in Rectifier (watts)	2.99
Forward Loss in Rectifier (watts)	3.20
Total Rectifier Loss (watts)	6.19
Loss per Plate (watts)	1.24
Temperature Rise (degrees centigrade)	43

radio receivers using a single rectifier. Fig. 9 is the circuit of a power supply using two rectifiers in a voltage-doubling arrangement. Fig. 10 shows a comparison of the voltage regulation of a 25Z6 tube and two 403D2625 selenium rectifiers, using voltage-doubling circuits in each case. It is to be noted that it is possible to use a much higher value of capacitance with the selenium rectifiers than with the tubes. This may be attributed to to the fact that the maximum permissible peak current for the tubes is less than that for the selenium rectifiers. Higher capacitances result in higher output voltages.

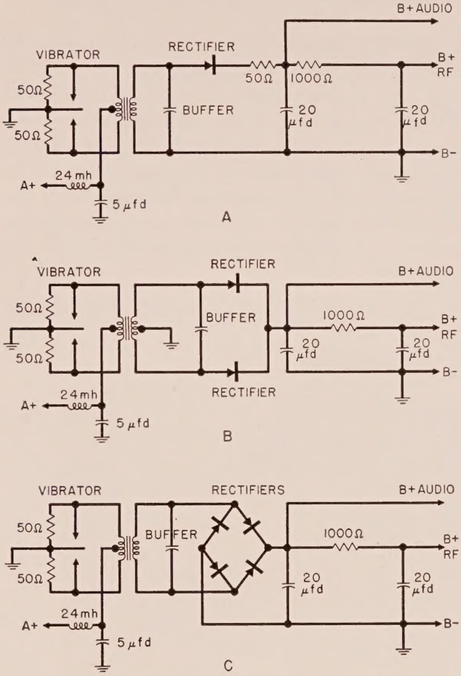

VIBRATOR
RECTIFIER
B+AUDIO
B+ RF
B−
50Ω
1000Ω
BUFFER
20 μfd
24 mh
A+
5 μfd
A
RECTIFIER
B
RECTIFIERS
C

Fig. 7—Three vibrator-type power supplies using selenium rectifiers in (A) half-wave, (B) full-wave, and (C) bridge-connected circuits, respectively.

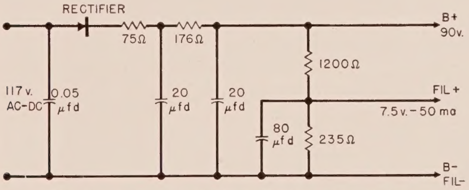

RECTIFIER
75Ω
176Ω
B+ 90v.
1200Ω
117 v. AC-DC
0.05 μfd
20 μfd
20 μfd
FIL+
7.5 v.−50 ma
80 μfd
235Ω
B− FIL−

Fig. 8—Power supply for AC/DC type receiver.

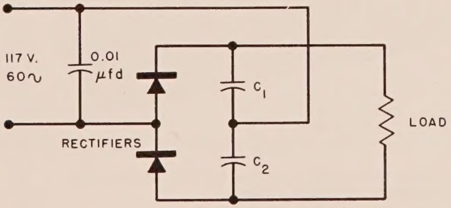

117 V. 60∿
0.01 μfd
C1
LOAD
RECTIFIERS
C2

Fig. 9—Selenium rectifiers in a voltage-doubling circuit.

One of the additional features in the use of the selenium rectifier is that no warm-up period is required for operation, as in case of the conventional tube-type rectifiers, making operation of the radio receiver practically instantaneous.

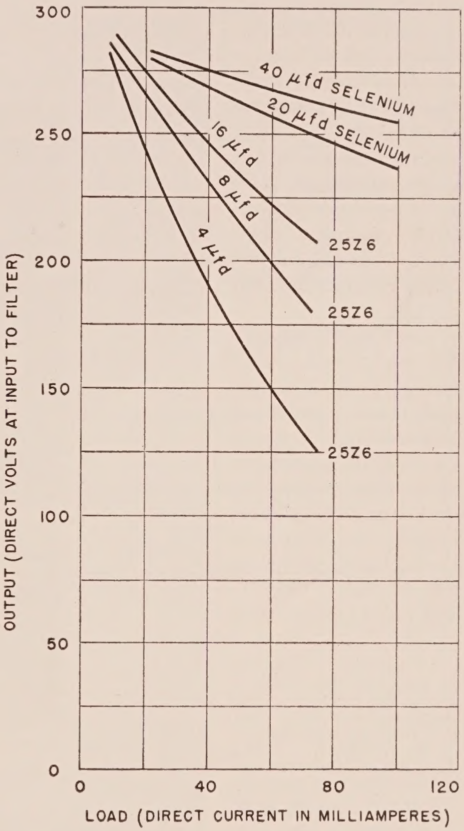


Fig. 10—Comparison of voltage regulation and output of a 25Z6 two-section vacuum diode and two 403D2625 selenium rectifiers. A much higher input capacitance to the filter may be used with the selenium rectifiers because of their greater permissible peak forward-current rating.

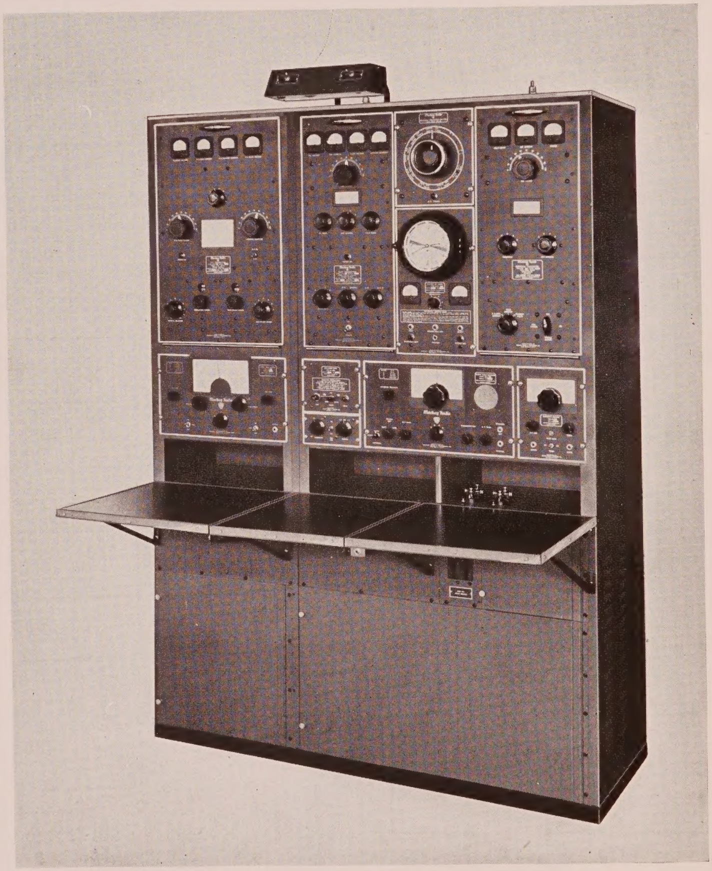
Single-Unit Radio Equipment for Passenger and Cargo Vessels

By W. J. GILLULE*

Mackay Radio and Telegraph Company, New York, New York

IN the prewar years, the fitting of the many separate units of radio equipment on shipboard, including hundreds of feet of connecting cable and wire, usually required several hundred man-hours. When the U. S. Maritime Commission, in 1940 and 1941, put into effect plans for building thousands of ships, Federal Telephone & Radio Corporation initiated the design of a complete marine radio unit which housed all the transmitter, receiver, and automatic alarm equipment in a single factory-wired assembly. Installation of this complete radiotelegraph station could usually be effected in one working day. This saved hundreds of thousands of skilled man-hours in building Liberty ships, Victory ships, and tankers.

The FT-101 and FT-105 marine radio units[1] incorporated a 300-watt main and emergency intermediate-frequency transmitter, automatic alarm, main and crystal-detector receivers, battery-charging controls, and antenna switch. Later, the FT-102 unit,[2] incorporating a 200-watt high-frequency transmitter and an all-wave receiver, was developed for installation on vessels equipped with FT-101 and FT-105 units to provide high-frequency communication facilities. In 1944, the FT-106 marine radio unit, combining all the apparatus of the FT-105 and FT-102 units, was developed for installation aboard the fleet of Victory ships and certain classes of tankers. The FT-106



Console-type marine radio equipment fullfilling all requirements for passenger and cargo vessels.

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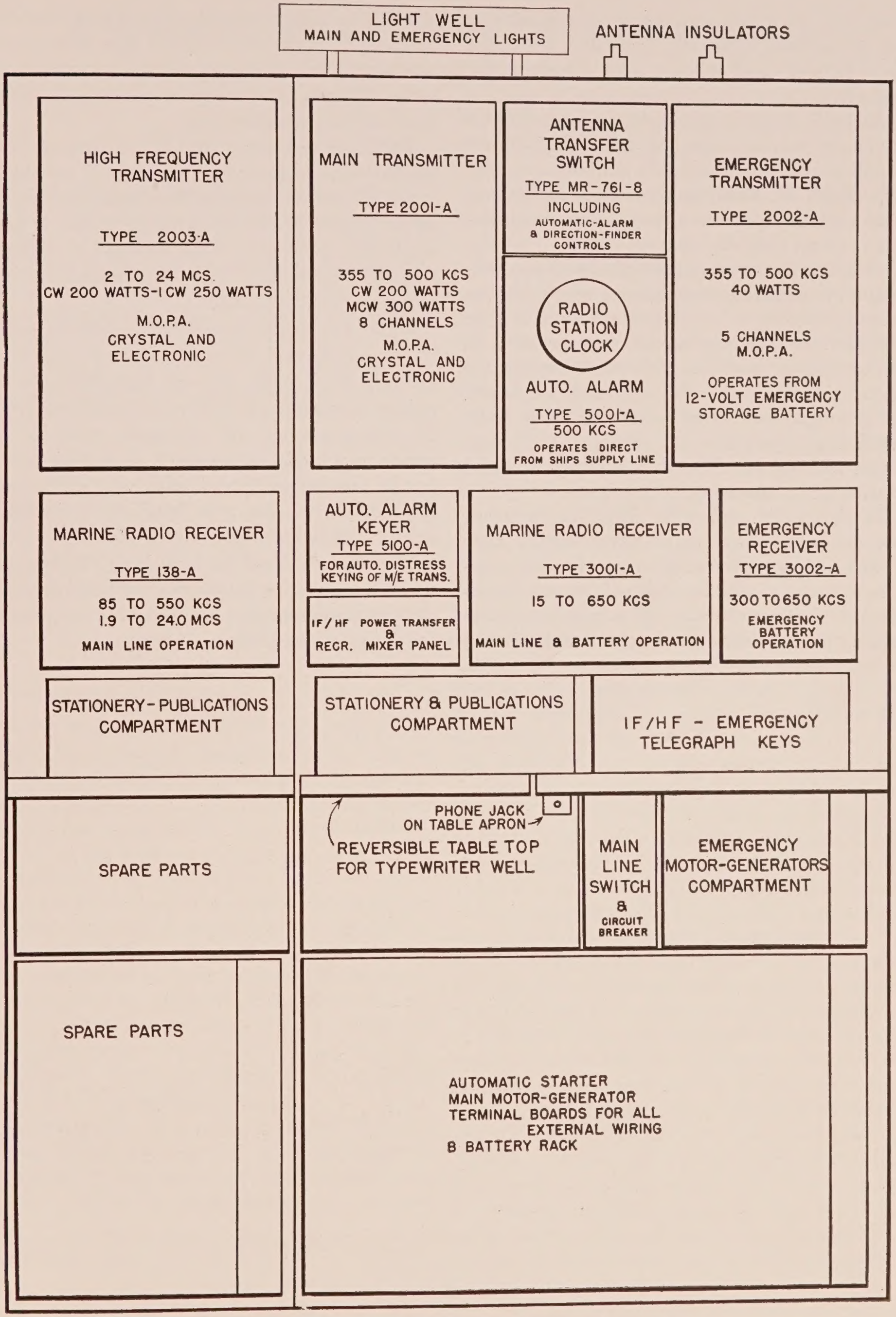
* The author was assisted in preparing this paper by E. G. Henry, who was then on the staff of Mackay Radio and Telegraph Company.

[1] E. J. Girard, "A New Marine Radio Unit for Cargo Vessels," *Electrical Communication*, v. 20, n. 2, pp. 71–72; 1941.

[2] E. J. Girard, "The H. F. Marine Radio Unit," *Electrical Communication*, v. 21, n. 2, pp. 85–88; 1943.


LIGHT WELL
MAIN AND EMERGENCY LIGHTS
ANTENNA INSULATORS
HIGH FREQUENCY TRANSMITTER
TYPE 2003-A
2 TO 24 MCS.
CW 200 WATTS-I CW 250 WATTS
M.O.P.A.
CRYSTAL AND ELECTRONIC
MAIN TRANSMITTER
TYPE 2001-A
355 TO 500 KCS
CW 200 WATTS
MCW 300 WATTS
8 CHANNELS
M.O.P.A.
CRYSTAL AND ELECTRONIC
ANTENNA TRANSFER SWITCH
TYPE MR-761-8
INCLUDING
AUTOMATIC-ALARM & DIRECTION-FINDER CONTROLS
RADIO STATION CLOCK
AUTO. ALARM
TYPE 5001-A
500 KCS
OPERATES DIRECT FROM SHIPS SUPPLY LINE
EMERGENCY TRANSMITTER
TYPE 2002-A
355 TO 500 KCS
40 WATTS
5 CHANNELS
M.O.P.A.
OPERATES FROM 12-VOLT EMERGENCY STORAGE BATTERY
MARINE RADIO RECEIVER
TYPE 138-A
85 TO 550 KCS
1.9 TO 24.0 MCS
MAIN LINE OPERATION
AUTO. ALARM KEYER
TYPE 5100-A
FOR AUTO. DISTRESS KEYING OF M/E TRANS.
IF/HF POWER TRANSFER & RECR. MIXER PANEL
MARINE RADIO RECEIVER
TYPE 3001-A
15 TO 650 KCS
MAIN LINE & BATTERY OPERATION
EMERGENCY RECEIVER
TYPE 3002-A
300 TO 650 KCS
EMERGENCY BATTERY OPERATION
STATIONERY-PUBLICATIONS COMPARTMENT
STATIONERY & PUBLICATIONS COMPARTMENT
IF/HF - EMERGENCY TELEGRAPH KEYS
SPARE PARTS
PHONE JACK ON TABLE APRON
REVERSIBLE TABLE TOP FOR TYPEWRITER WELL
MAIN LINE SWITCH & CIRCUIT BREAKER
EMERGENCY MOTOR-GENERATORS COMPARTMENT
SPARE PARTS
AUTOMATIC STARTER
MAIN MOTOR-GENERATOR
TERMINAL BOARDS FOR ALL EXTERNAL WIRING
& BATTERY RACK

Layout of type MRU-10/11 equipment for passenger and cargo vessels.

measured 87 inches wide by 69 inches high by 19 inches deep (less the operating shelf and light fixtures) and weighed approximately 1800 pounds.

In 1945, Mackay developed the intermediate-frequency marine radio unit type MRU-10 and a companion high-frequency marine radio unit type MRU-11. The objective was to make them even more compact than the FT-106, and introduce certain desirable improvements.

One of the major improvements is the provision of completely separate emergency equipment consisting of emergency transmitter type 2002-A and emergency receiver type 3002-A, both operating from a 12-volt emergency storage battery. This provision is particularly important as the equipment then conforms with radio regulations applicable to passenger and cargo ships, whereas the FT-106 could only be fitted aboard cargo vessels.

Marine radio unit type MRU-10/11 incorporates a new automatic-alarm keyer, type 5100-A, which operates from the 12-volt emergency battery and is designed to key automatically either the main or emergency transmitter for sending the international distress signal. The new automatic alarm, type 5001-A, operates from the ship's supply line, thus eliminating two sets of 24-volt storage batteries used in the FT-106, an appreciable saving in installation and maintenance costs.

The main receiver, type 3001-A, superseding the battery-operated receiver, 128-AV, provided in the FT-106, is energized from either the ship's power supply or, in the event of failure of the ship's line, from the 12-volt emergency battery and a 90-volt B battery. Power transfer is accomplished by a single knob.

While incorporating several additional units of equipment and numerous features not provided in the FT-106, the new type MRU-10/11 weighs approximately 1200 pounds and the overall dimensions are only 54 inches wide by 74 inches high by 19 inches deep. It is 43.7 cubic feet in volume compared with 65.6 cubic feet for the FT-106, the type MRU-10/11 being $\frac{1}{3}$ smaller and lighter than the FT-106.

The appearance of the type MRU-10/11, and an outline drawing indicating the relative position of the component equipments and general specifications are shown.

A Method of Measuring Grid Primary Emission in Thermionic Valves*

By A. H. HOOKE

Valve Division, Standard Telephones and Cables, Ltd., London, England

THE paper describes equipment designed to measure grid primary emission in thermionic valves under controlled conditions of grid and anode dissipation. An explanation of grid primary emission and an outline of some of the difficulties associated with its measurement are given. These are followed by a description of the circuit of the measuring instrument. The way in which power is applied to the grid and anode of the valve under test allows the operating conditions of the valve to be adjusted as desired. By taking measurements on a valve with first, zero anode dissipation, then any conveniently selected value of anode dissipation, the grid dissipation necessary to produce a given emission current may be calculated for any value of anode dissipation. Examples of measurements, which indicate the close agreement between measured and calculated values, are given.

• • •

1 *Introduction*

Apart from those electronic valves in which certain electrodes are designed with a view to utilizing their properties as emitters of secondary electrons, emission from any electrode other than the cathode can generally be regarded as detrimental to the efficient performance of a valve. An electrode will emit primary electrons when its temperature is raised sufficiently to overcome the work function of the material of which it is made. The work functions of materials used in the construction of most valve electrodes other than cathodes are high, and if the electrodes are maintained in an uncontaminated condition, it is improbable in most practical cases that the temperature would rise sufficiently for emission to occur to any significant degree.

Unfortunately, it is often extremely difficult during certain processes in valve manufacture involving the generation of considerable temperature within the valve structure, to prevent contamination of electrode surfaces by active materials emanating from such sources as the cathode.

Owing to its proximity to the cathode the control grid in the average valve is often the most subject to contamination, and for the same reason it may tend, under operating conditions, to rise in temperature due to radiant heat from the cathode surface.

Transmitting valves due to their high operating temperatures, are more subject to grid primary emission than are receiving valves although it is by no means unknown among the latter particularly in closely spaced types.

The determination of grid primary emission is thus a problem in which both the valve designer and the circuit engineer are directly interested. Unfortunately the direct measurement of grid primary emission presents some inherent difficulties. The temperature reached by the control grid is often substantially affected by back-heating from other electrodes such as the anode as well as from the cathode. This back-heating will obviously be a function of the power dissipated in the other electrodes so that grid primary emission may be expected to be a maximum under conditions of maximum power dissipation on all electrodes. Unfortunately this condition of maximum power dissipation may be conducive to the release of gas in the envelope, also sometimes due to bombardment by high-velocity electrons from the cathode, the grid may emit secondary electrons. These two effects may both result in grid current in the same direction and possibly of the same order of magnitude as the current resulting from primary emission. In normal methods of measurement the grid currents are measured collectively and no discrimination between the amounts due to the three components (primary emission and the two components just mentioned above) is possible. Also when the grid is swinging sufficiently positive, the mean resultant grid current will be in a direction opposite to that of the primary and secondary emission

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* Reprinted from *Electronic Engineering*, v. 18, pp. 75–80; March, 1946.

currents so that these will be masked from measurement by normal means. Hence, in order to obtain adequate data regarding the behavior of grid primary emission it is necessary to have available a device capable of discrimination between this and the other two components mentioned above. As the equipment described in this paper is designed to measure grid primary emission, all following references to grid emission should be understood to apply to primary emission only unless otherwise stated.

With the apparatus to be described, measurements of grid emission may be made at any desired grid dissipation without danger of damage to the valve. In addition facilities are provided for determining in triode valves the effect on grid emission of back-heating from the anode. Determination of the back-heating effects in tetrodes and pentodes presents difficulties owing to complications that arise, both in the circuit of the instrument and in the calculation of electrode dissipations other than that of the grid. The instrument does not, therefore, permit of measurements on such valves under normal operating conditions but adequate data can, as a rule, be obtained by operating these valves as triodes since heating of the control grid from sources other than the cathode and anode can usually be considered negligible.

The circuit of the instrument is so arranged that the anode and grid of the valve under test are consecutively supplied with heating power followed by a measurement of any grid emission present, these operations being repeated in a cyclic order at a constant repetition frequency. The grid emission is indicated on a suitable meter as an average of the current flowing during the period occupied by an operational cycle. In any one cycle of operation, power is supplied twice to the grid and anode followed by the application to the grid of a high negative voltage sufficient to cut off all anode current. It is during this last period that the grid primary emission is measured.

The heating powers for both anode and grid are derived from the 50-cycle-per-second mains supply through half-wave rectifiers, so that half-cycles of alternating voltage only are delivered to the valve. A phase difference of 180 degrees exists between the anode and grid supplies thus allowing the anode voltage to be applied during the periods of one half-cycle duration, in which the grid has zero potential. Meters are included in each circuit so that the power input to each electrode may be calculated.

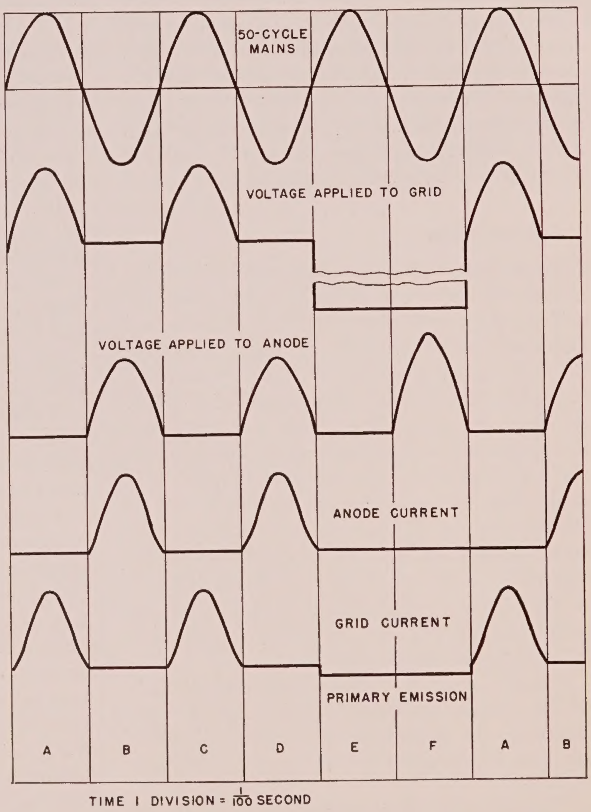

Fig. 1—Relative phases of anode and grid currents.

The complete sequence of operations may be seen by reference to Fig. 1, which indicates the relative phases of the anode and grid voltages and currents, together with a 50-cycle-per-second wave for comparison purposes. For the purpose of explanation, the operational cycle has been divided into six periods of 1/100th second (half-cycle intervals at 50 cycles per second) which have been labelled *A*, *B*, *C*, *D*, *E*, and *F*. The amplitudes of the various waves shown in Fig. 1 are drawn to an arbitrary scale and are not significant.

To start the analysis of the sequence, let us consider the interval *A*. During this period the grid of the valve under test is supplied with a half-cycle voltage wave of positive polarity, causing grid current to flow and power to be dissipated, while the anode potential is zero. In the interval *B* which follows, it is the anode which has a half-cycle wave applied to it, with the grid remaining at zero potential. Interval *C* is a repetition of *A* and likewise interval *D* is the same as *B*. During the intervals *E* and *F* a constant high negative voltage is applied to the grid. The anode is supplied with zero voltage during *E* and a positive half-cycle during *F*, but the negative voltage on the grid is of sufficient magnitude to carry the valve well past anode-current cut-off during all of both intervals *E* and *F*. Interval *F* concludes the sequence, following intervals being repetitions of *A*, *B*, *C*, *D*, *E*, and *F*.

It will be seen that during the period *A*, *B*, *C*, *D*, *E*, *F*, the grid is supplied with power for two of the 1/100-second intervals and so is heated for one-third of the period occupied by an operational cycle. If the grid attains a high enough temperature, grid emission will occur and the resulting current may be measured in the external circuit during the periods *E* and *F*.

It will be noticed that the voltage applied to the anode during interval *F* is shown as being greater than that in intervals *B* and *D*. This is due to the absence of load on the power supply unit and cannot be avoided easily. However, provision has been made to prevent this voltage from affecting the reading of anode dissipation by automatically disconnecting the anode voltmeter during interval *F*.

Referring to Fig. 1 again, it will be seen that the currents corresponding to the voltages applied to the anode and grid are also indicated in the appropriate intervals. As the instantaneous anode currents are proportional to the instantaneous anode voltages raised to the power of 3/2, the current pulses, as seen in the figure, will not have the shape of half sine waves.

2 *Circuit and Operation of Instrument*

The circuit diagram is shown in Fig. 2. The operation is as follows: A 50-cycle-per-second supply is fed via a phase adjuster and transformer into a half-wave rectifier *V*11. The output from this rectifier, developed across its load resistance *R*13 in the form of a series of half-cycles, is fed to the control electrodes of three gas-filled, cold-cathode, three-electrode valves arranged in a ring counting circuit *V*6, *V*7, and *V*8. One of these valves ionises every time an impulse is received and deionises when the next in the ring is ionised by the succeeding impulse. Hence, any one valve strikes every third impulse received, which in this case is every third half-cycle swing across the load resistance *R*13, corresponding to every third cycle of the mains supply. One of the valves of the ring is provided with a load resistance *R*3 in its anode circuit for the purpose of feeding a synchronising signal to a multivibrator *V*9.

The multivibrator provides an asymmetrical output of rectangular form at a frequency of 50/3 cycles per second and is locked to the main 50-cycle-per-second supply in the manner described above. The width of the negative pulse generated by this multivibrator is half that of the positive pulse, the duration of their sum being that of three cycles of mains frequency.

The output from the multivibrator is applied to the grid of a biased amplifier valve *V*10 which, in turn, controls two relays, *K*1 and *K*2. *K*1 switches the grid of the valve under test *Vt* to its appropriate supplies and *K*2 switches the anode voltmeter of *Vt*.

When the multivibrator delivers a positive pulse to the grid of the biased amplifier, current will flow in the anode circuit causing *K*1 and *K*2 to operate and to remain operated for a period corresponding to the width of the positive pulse. In the operated position *K*1 connects the grid of the valve under test *Vt* to the source of heating power via the rectifier *V*1 and the transformer *T*1.

In the unoperated position *K*1 connects the grid of *Vt* to a high negative voltage via a grid current meter which measures the grid primary emission.

Thus when $K1$ is operated (for a period corresponding to the positive pulse width of the multivibrator) the grid of Vt is heated by half sine-wave pulses of power from $V1$. $K1$ then releases and, with Vt biased well past anode-current cut-off by $P1$, any grid primary emission is measured by the grid current meter. When discussing the multivibrator it was seen that the positive pulse lasted for 1/25 second. This corresponds to intervals A to D on Fig. 1 where it is seen that during this time the grid is supplied with power for a total of two 1/100-second intervals; for the other two 1/100-second intervals the grid is at zero potential and it is during these latter intervals that power may be supplied to the anode from a source similar to that used for the grid. Intervals E and F correspond to the period of the negative pulse of the multivibrator during which $K1$ and $K2$ are unoperated and primary grid emission is being measured on the grid microammeter. This meter will indicate an average of the current flowing during the periods A to F.

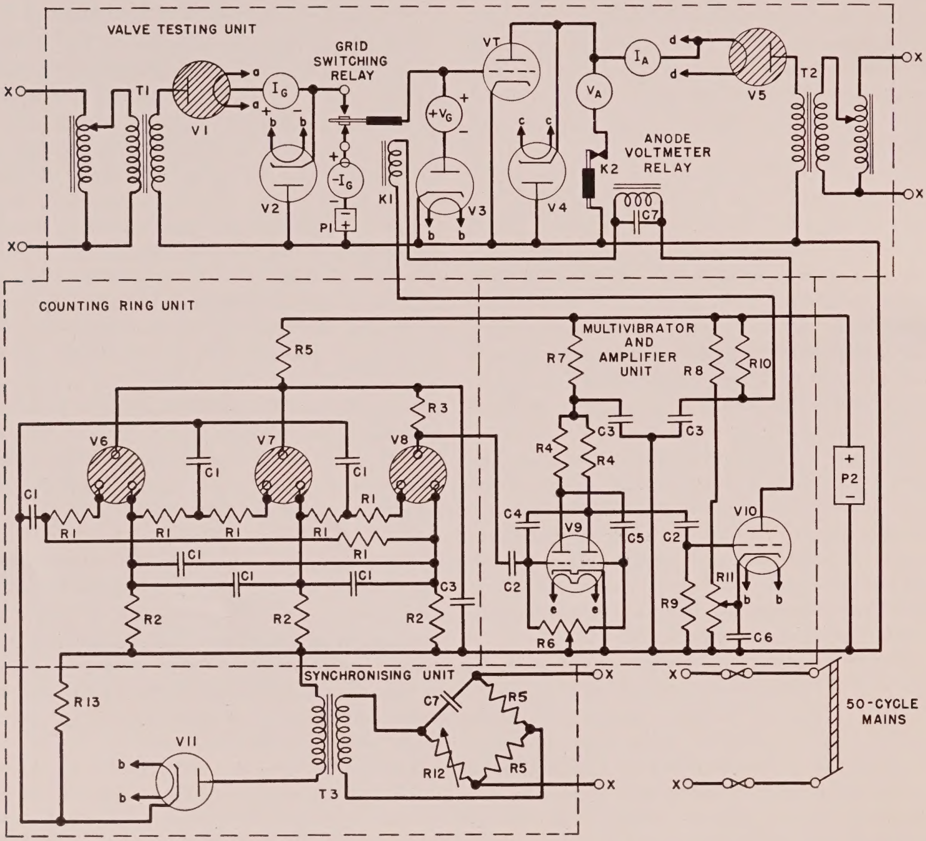
VALVE TESTING UNIT
GRID SWITCHING RELAY
X
T1
V1
a
a
I_G
b
b
V2
+
$-I_G$
P1
K1
$+V_G$
V3
VT
c
c
V4
I_A
V_A
d
d
K2
ANODE VOLTMETER RELAY
C7
V5
T2
X
X
COUNTING RING UNIT
MULTIVIBRATOR AND AMPLIFIER UNIT
R5
R3
R7
R8
R10
V6
V7
V8
C1
R1
R2
C3
R4
C4
V9
C5
C2
e
e
R6
R9
R11
V10
C6
P2
SYNCHRONISING UNIT
R13
V11
T3
C7
R5
R12
R5
50-CYCLE MAINS

Fig. 2—Circuit diagram of measuring apparatus.

The purpose of the anode voltmeter is, in conjunction with the anode current meter, to measure the power supplied to the anode of the valve under test. During intervals *E* and *F* no anode current flows in *Vt* owing to the large negative grid voltage and so no power is delivered to the anode during this period. The relay *K*2 is therefore inserted to cut the anode voltmeter out of circuit during intervals *E* and *F*. If this were not done the high peak no-load voltage applied during interval *F* would unnecessarily complicate the interpretation of the voltmeter reading in calculating the heating power applied to the anode.

It can be appreciated that the relays are working under ideal conditions, since they are synchronised with the anode and grid supplies so that there is no current flowing across the contacts during their operation.

The anode and grid half-wave power supplies are of conventional design and require no explanation.

Synchronisation of the instrument to the mains is effected by means of the phase adjuster at the input. This phase adjuster, being of the single-reactance bridge type delivering a constant voltage output, is very stable and once set rarely requires adjustment.

As an alternative to the method described for the co-ordination of the supplies to the valve under test, a synchronous motor, driving suitable commutators, could be employed. A later model of this equipment has, in fact, been modified in this manner.

3 *Interpretation of Meter Readings*

The meters employed throughout the instrument are of the direct-current moving-coil type and therefore register the average values of the wave-forms applied. The desired root-mean-square values may, therefore, be obtained by the use of appropriate multiplying factors.

The use of a rectangular waveform for operating the grid switching relay *K*1 simplifies the calculation of the actual grid emission current. Since the relay is on its back contact for one-third of each cycle of operation (*A* to *F* in Fig. 1), the actual emission current meter reading is one-third the true value.

The interpretation of the grid supply meter readings is somewhat different. Here we are interested in the input power rather than the root-mean-square values corresponding to the readings of the individual meters. Hence it is more desirable to apply a single multiplying factor to the product of the grid input voltmeter and ammeter readings in order to arrive at the figure for the average power dissipation. It has been found that the multiplying factor required for this operation is 3.8 (Appendix I).

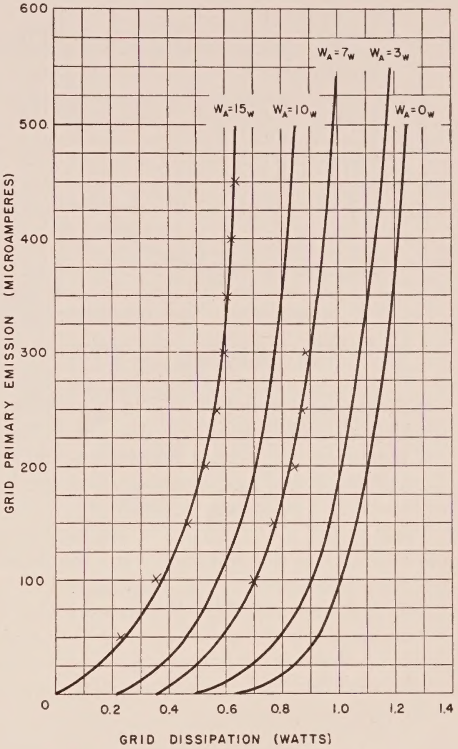

Fig. 3—Grid emission and dissipation for low-power-output valve.

As the anode voltmeter is disconnected by *K*2 during all *F* intervals (Fig. 1), the waveforms applied to this meter and to the anode current meter are the same as those applied to the corresponding grid supply meters, so that the multiplying factor is the same as before, i.e., 3.8.

4 *Measurements*

The measurement of grid primary emission on valves so far carried out with this instrument has been directed towards the production of curves showing the effect of varying the grid dissipation for various values of anode dissipation. Complete sets of curves can be taken in this manner by setting up the valve under test in the instrument, selecting the required anode dissipation, and taking readings for varying grid voltage values. The curves obtained are consistent and can be repeated with considerable accuracy.

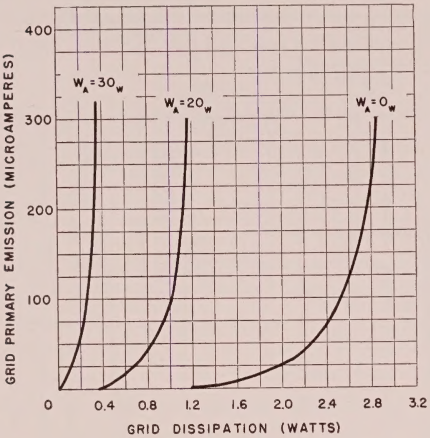

Fig. 4—Effect of anode power on a close-spaced power valve.

When a full set of curves is required, it is not necessary, however, to follow this somewhat laborious procedure. It has been argued in a paper by Mouromtseff and Kozanowski[1] that if the assumption is made that all the power dissipated by the grid structure is radiated to the anode, then for any given grid emission, that is to say, grid temperature:

$$W_g = A - kW_a, \tag{13}$$

where A and k are constants, and W_g and W_a represent the energies dissipated at the grid and anode, respectively (Appendix II).

The use of this principle allows of an appreciable saving in time and labour, as measurements need only be taken for the plotting of two curves of grid emission versus grid dissipation, and the second with any convenient anode dissipation. From this information experimentally derived, it is then possible to calculate, for any value of grid emission current under consideration, the necessary grid input power required at any value of anode dissipation.

The above procedure has been utilized to produce many curves, some of which have been experimentally verified to provide a check. In these it has been found that the calculated and measured values show close agreement, a maximum divergence of 5 per cent, with a usual divergence of 2 per cent, from measured values being observed.

5 Conclusion

The curves taken with this instrument have provided information which has proved of great value, particularly in experiments upon valves involving different materials and methods of construction of the grid assembly. Four typical examples of such curves are shown in Figs. 3–5.

Fig. 3 represents a set of curves taken for a low-power-output valve. All the drawn curves are experimental and the crosses represent calculated values using the experimental results for zero and 10 watts anode dissipation as a basis. The close agreement between measured and calculated values is apparent.

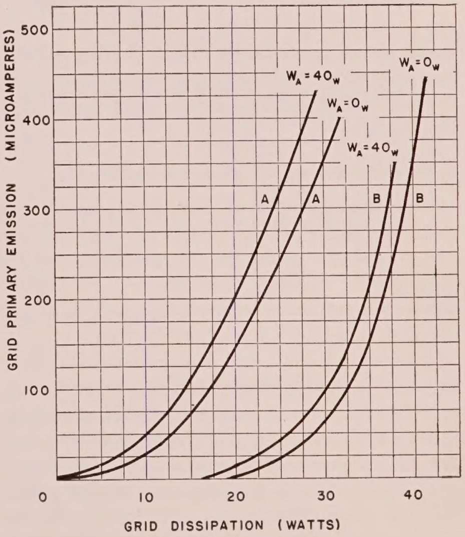

Fig. 5—Effect of grid material on emission characteristics. *AA* = valve with molybdenum grid, *BB* = valve with tantalum grid.

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[1] Mouromtseff and Kozanowski, "Grid Temperatures as a Limiting Factor in Vacuum Tube Operation," *Proceedings of the I.R.E.*, v. 24, pp. 447–454; March, 1936.

Fig. 4 gives an indication of the effect of anode power upon a very close-spaced power valve. It is to be noted that for a given grid dissipation the grid emission increases very rapidly with increase of anode power.

Fig. 5 indicates the effect of variation in grid material upon primary grid emission characteristics. It will be noted that the valve employing a tantalum grid will stand appreciably more input to that electrode for a given emission current than will a similar valve in which the grid was constructed from molybdenum.

6 Appendices

6.1 DETERMINATION OF MULTIPLYING FACTOR FOR THE CONVERSION OF THE INDICATED VOLTAGES AND CURRENTS INTO MEAN POWERS

Let I_{av} = measured average current and E_{av} = measured average voltage. Let the load present a linear impedance:

Then the average current in one half-cycle $=\frac{2}{\pi} I_{peak}$.

Hence by reference to Fig. 1, measured average current over six half-cycles $= I_{av}$

$$=\frac{2}{6}\times\frac{2}{\pi}I_{peak}=\frac{2}{3\pi}I_{peak}. \tag{1}$$

Now root-mean-square current in one half-cycle

$$=\frac{1}{\sqrt{2}}I_{peak}=\frac{1}{\sqrt{2}}\times\frac{3\pi}{2}I_{av}.$$

Also it follows from this that root-mean-square voltage in one half-cycle

$$=\frac{1}{\sqrt{2}}\times\frac{3\pi}{2}E_{av},$$

∴ mean power in one half-cycle

$$=\left(\frac{1}{\sqrt{2}}\times\frac{3\pi}{2}\right)^2 E_{av}I_{av}$$

and power in six half-cycles

$$=\frac{2}{6}\left(\frac{1}{2}\times\frac{3\pi}{2}\right)^2 E_{av}I_{av}=3.7E_{av}I_{av}. \tag{2}$$

However, the current in a valve does not obey a linear law but is usually a function of the three-halves power law over most of the operating range.

Let i = instantaneous value of current, I_m = peak current, and E_m = peak voltage.

Then $i=kE^{3/2}$.

Now

$$\frac{i}{I_m}=\frac{(E_m\cos\theta)^{3/2}}{E_m{}^{3/2}}=\cos^{3/2}\theta,$$

$$\therefore\quad i=I_m\cos^{3/2}\theta.$$

Hence mean power over quarter-cycle

$$=\frac{I}{\pi/2}\int_0^{\pi/2} I_m\cos^{3/2}\theta\times E_m\cos\theta\,d^{\theta}$$

$$=\frac{I_mE_m}{\pi/2}\int_0^{\pi/2}\cos^{5/2}\theta\,d^{\theta}. \tag{3}$$

Now it is known that

$$\int_0^{\pi/2}\cos^{m-1}x\sin^{n-1}x\,dx=\tfrac{1}{2}\frac{\gamma(\frac{1}{2}m)\gamma(\frac{1}{2}n)}{\gamma(\frac{1}{2}m+\frac{1}{2}n)}, \tag{4}$$

which is the same as equation (3) if

$$x=\theta,\quad n=1\quad\text{and}\quad m=(1+5/2)=7/2.$$

Hence mean power in quarter-cycle

$$=\frac{I_mE_m}{\pi/2}\left[\frac{1}{2}\frac{\gamma(\frac{1}{2}\times 7/2)\gamma(\frac{1}{2}\times 1)}{\gamma(\frac{1}{2}\times 7/2+\frac{1}{2}\times 1)}\right]$$

$$=\frac{I_mE_m}{\pi/2}\left[\frac{1}{2}\frac{\gamma(7/4)\gamma(\frac{1}{2})}{\gamma(9/4)}\right].$$

From table of gamma functions

$$\gamma\left(\frac{7}{4}\right)=\gamma(1.75)=0.9191,$$

$$\gamma\left(\frac{1}{2}\right)=\sqrt{\pi},$$

$$\gamma\left(\frac{9}{4}\right)=\left(\frac{9}{4}-1\right)\gamma\left(\frac{9}{4}-1\right)=\frac{5}{4}\times 0.9064,$$

∴ mean power in quarter-cycle

$$=\frac{I_mE_m}{\pi/2}\left(\frac{1}{2}\times\frac{0.9191\times\sqrt{\pi}}{5/4\times 0.9064}\right),$$

$$=\frac{0.7190I_mE_m}{\pi/2}$$

and mean power in one half-cycle

$$=2\times 0.7190\,\frac{I_mE_m}{\pi}=0.4576\,\frac{I_mE_m}{\pi}.$$

Now

$$i=I_m\cos^{3/2}\theta, \tag{5}$$

∴ average current in one quarter-cycle

$$=\frac{I}{\pi/2}\int_0^{\pi/2} I_m \cos^{3/2}\theta\, d\theta,$$

$$=\frac{I_m}{\pi/2}\int_0^{\pi/2} \cos^{3/2}\theta\, d\theta,$$

$$=\int_0^{\pi/2} \cos^{m-1}\times\sin^{n-1}\times dx.$$

If

$$n=1,\quad m=(1+3/2)=5/2 \quad\text{and}\quad x=\theta,$$

∴ average current in one quarter-cycle

$$=\frac{I_m}{\pi/2}\left[\frac{1}{2}\,\frac{\gamma(\tfrac{1}{2}\times 5/2)\gamma(\tfrac{1}{2}\times 1)}{\gamma(\tfrac{1}{2}\times 5/2+\tfrac{1}{2}\times 1)}\right],$$

$$=\frac{I_m}{\pi/2}\left[\frac{1}{2}\,\frac{\gamma(5/4)\gamma(\tfrac{1}{2})}{\gamma(7/4)}\right]$$

and from table of gamma functions

$$\gamma(5/4)=0.9064,\quad \gamma(1/2)=\sqrt{\pi},\quad \gamma(7/4)=0.9191,$$

∴ average current in one quarter-cycle

$$=\frac{I_m}{\pi/2}\left[\frac{1}{2}\times\frac{0.9064\times\sqrt{\pi}}{0.9191}\right]=0.8738\frac{I_m}{\pi/2}$$

and average current over one half-cycle is obviously the same

$$=\frac{2\times 0.8738 I_m}{\pi}=0.5560 I_m. \tag{6}$$

But measured average current over period of six half-cycles

$$=I_{av}=\frac{2}{6}\times 0.5560 I_m=0.1853 I_m. \tag{7}$$

Now we have proved in (5) that mean power in half-cycle

$$=0.4576 E_m I_m,$$

then for I_m we can substitute $\frac{I_{av}}{0.1853}$ and for E_m we can write $\frac{3\pi}{2}E_{av}$ similar to (1),

∴ mean power in half-cycle

$$=0.4576\times\frac{3\pi}{2}E_{av}\times\frac{I_{av}}{0.1853}.$$

Hence mean power over six half-cycles

$$=\frac{2}{6}\times\frac{0.4576\times 3\pi}{2\times 0.1853}E_{av}I_{av}$$

$$=3.88 E_{av}I_{av}, \tag{8}$$

which is higher than the value obtained for a linear law. However, it is realized that the above value would not be so large in practice due to electron scattering, etc., and a figure of $3.8\times E_{av}I_{av}$ has been taken to represent the mean power.

6.2 The Relationship between Anode and Grid Dissipations for a Given Grid Primary Emission

Assuming that the total power dissipated in the grid is radiated to the anode, let W_g = grid dissipation, and W_a = anode dissipation, then

$$W_g=c(T_g^4-T_a^4), \tag{9}$$

where c is a constant for a given valve structure, T_g and T_a are grid and anode temperatures.

A similar relationship exists for the radiation from the anode into the surrounding space

$$W_a+W_g=C_1(T_a^4-T_0^4), \tag{10}$$

where T_0 is the temperature of the surrounding medium. We may write:

$$\frac{W_a+W_g}{C_1}=T_a^4-T_0^4,$$

$$\therefore\quad T_a=\frac{W_a+W_g}{C_1}+T_0^4. \tag{11}$$

Substituting for T_a^4 in (9)

$$W_g=C\left(T_g^4-\frac{W_a+W_g}{C_1}+T_0^4\right),$$

$$=\frac{CW_a+CW_g}{C_1}+CT_g^4+CT_0^4,$$

$$\therefore\quad W_g+\frac{CW_a+CW_g}{C_1}=CT_g^4+CT_0^4$$

and

$$W_g\frac{(C_1+C)}{C_1}=(CT_g^4-CT_0^4)-\frac{CW_a}{C_1},$$

$$W_g=\frac{(CT_g^4-CT_0^4)C_1}{C_1+C}-\frac{CW_a}{(C_1+C)}, \tag{12}$$

then

$$\frac{dW_g}{dW_a}=\frac{-C}{(C_1+C)}$$

or the rate of increase of W_g with respect to W_a is constant, which gives a linear relationship,

$$\therefore\quad \text{equation}\qquad W_g=A-kW_a \tag{13}$$

may be used.

Recent Telecommunication Developments

Federal Electric Manufacturing Company, Ltd.—A newly organized Canadian subsidiary of Federal Telephone and Radio Corporation, which will be known as the Federal Electric Manufacturing Company, Ltd., has acquired a manufacturing plant in Canada for telephone, radio, and other electrical equipment.

A modern industrial structure has been purchased from the Canadian War Assets Corporation. Containing approximately 90 000 square feet of floor space, it is situated about five miles from the center of Montreal. An additional eight acres of land adjoining the factory have been acquired for future expansion. Initial operations will employ approximately 200 Canadians.

Available to the Canadian organization will be the research, development, and engineering resources of the entire I.T.&T. System which, through the International Standard Electric Corporation, co-ordinates the activities of 27 affiliated manufacturing companies in Great Britain, Continental Europe, Latin America, Australia, New Zealand, and China. Sales and service branches are equipped to distribute telecommunication, electronic, and electrical equipment in 75 countries.

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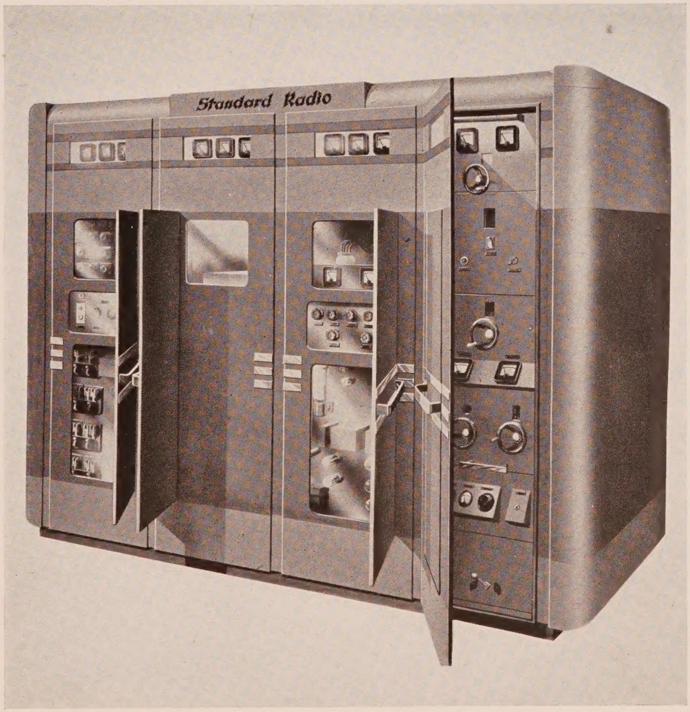
High-Frequency Transmitter—A new 5-kilowatt high-frequency radio transmitter, type D.S.10, has been developed by Standard Telephones and Cables, Ltd. This transmitter is capable of operating simultaneously on three continuous-wave channels, or on one continuous-wave and one telephone channel, and is continuously tunable from 2.2 to 22 megacycles per second. To simplify adjustment, the tuning controls of the exciter stages are ganged and the dials are calibrated directly in frequency.



The minimum assembly comprises two cabinets, containing a power unit and one radio-frequency unit. The dimensions are 5 feet 9 inches high, 5 feet 6 inches wide, and 2 feet 9 inches deep. Extra radio-frequency units up to a total of six, with a modulator, may be added if required.

The use of cathode-follower coupling in the class-B output stage of the modulator gives improved matching to the grids, and negative feedback is introduced to minimize distortion in the stage. A specially developed 5-kilowatt air-blast-cooled pentode valve is used in the radio-frequency output stage, thus simplifying tuning by eliminating the need for neutralization. The valve has a thoriated-tungsten filament, and a maximum plate-dissipation rating of 2.5 kilowatts. A high-voltage supply of 6000 volts is provided for continuous-wave operation and is reduced to 4500 volts for telephony. The size of the radio-frequency cabinet is kept small by the use of a single-ended output circuit.

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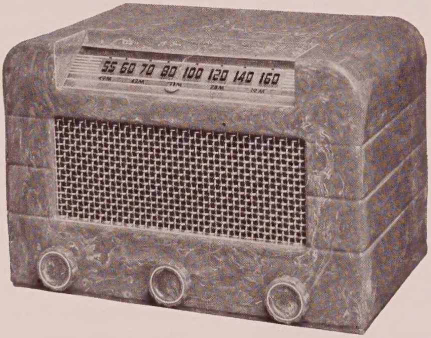
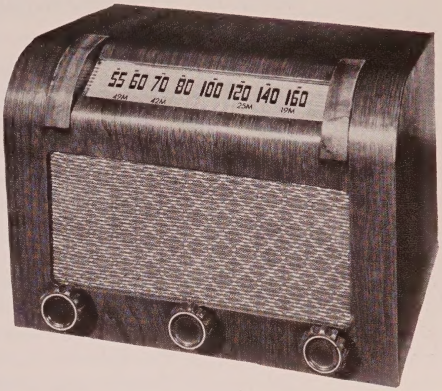
Broadcast Receivers—Federal Telephone and Radio Corporation is producing the initial models of household broadcast receivers of the AC/DC table type. A full series of console designs will follow.



Export design for 117- or 235-volt AC/DC operation.

Three of the receivers in production are electrically identical but are mounted in different types of wooden cabinets. Utilizing six vacuum tubes, they cover the standard U.S.A. broadcast frequencies and an additional band from 6 to 16 megacycles per second. A built-in loop makes an external antenna unnecessary for broadcast reception.



AC/DC 6-tube broadcast receiver for U.S.A. market.

A 5-tube receiver, covering the broadcast band and the high frequencies from 6 to 18 megacycles, is designed primarily for export. It may be connected either to 117- or 235-volt alternating- or direct-current circuits. For 117-volt operation, the ballast tube is inserted in the socket in a different position than for 235 volts, thus disconnecting the resistive elements required at the higher voltage. Molded bakelite cabinets are available in a variety of colors.

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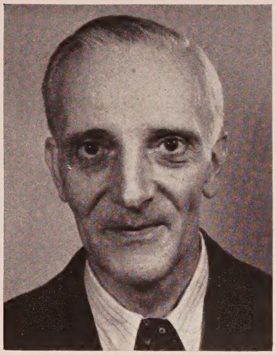
Telephoto Facsimile Equipment—An arrangement has recently been made whereby International Standard Electric Corporation will distribute in countries outside of the U.S.A. the telephoto facsimile equipment manufactured by Times Telephoto Equipment Inc., a subsidiary of the New York Times Company. This system, originally developed for the transmission of news photographs, was widely used for many purposes in the communication services during the war by the armed forces and the Office of War Information. The entire output has, until recently, been applied to the needs of the government and The Associated Press. Now that equipment is being produced in larger quantities, it will be possible to make it available to others throughout the world.

Contributors to This Issue



Leon N. Brillouin

Leon N. Brillouin was born on August 7, 1889, in Sevres, France. He was educated at the University of Paris, receiving a doctorate in 1920.

During World War I, he served in the French Army as a radio engineer. In 1928, he was appointed professor of theoretical physics at the Institut H. Poincaré of the University of Paris, and in 1932 joined the faculty of the College de France in the same capacity.

He served as general director of the French National Broadcasting System from July, 1939, to January, 1941, when he came to the U.S.A. From 1941 to 1942, he was a professor at the University of Wisconsin, and spent the next academic year on the staff of Brown University. He was engaged in defense research work at Columbia University from 1942 to 1945, and is now teaching at Harvard University.

Dr. Brillouin has been a consultant to Le Matériel Téléphonique in Paris since 1936, and to Federal Telecommunication Laboratories since 1941.

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Marie Bruzau

Marie Bruzau was born on January 4, 1900, in La Queue-en-Brie (Seine et Oise), France.

In 1922 she received the degree of engineer at the Ecole de Physique et de Chimie Industrielle of Paris, and in 1933 the degree of Docteur ès-Sciences Physiques at the Faculté des Sciences of Paris.

As chief of practical experimenting at the Ecole de Physique et de Chimie, Madame Bruzau did research work in organic chemistry and, in 1938, at the request of Laboratoire Central de Télécommunications, started a study of piezoelectric substances.

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Edward W. Chadwick

Edward W. Chadwick received a B.S. degree in electrical engineering from North Carolina State College in 1926. On leaving college, he was employed by the Brooklyn Edison Company on development and test work.

In 1942 he joined the inspection staff of the selenium rectifier department of the International Telephone and Radio Corporation which subsequently became the Federal Telephone and Radio Corporation. In 1944, he was transferred to the rectifier engineering department, and the next year became supervisor of the rectifier stack engineering department.

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A. G. Clavier

A. G. Clavier was born at Cambrai, France, in 1894. He received a degree in electrical engineering from Ecole Supérieure d'Electricité in 1919 and then joined the staff of engineers organized by General Ferrié at the Etablissement Central de la Radiotélégraphie Militaire. He was in charge of research on high frequencies from 1920 to 1925.

In 1929, Mr. Clavier joined Les Laboratoires Standards in Paris which later became Laboratoire Central de Télécommunications, and has been continuously engaged in research on centimeter and millimeter waves. He was in charge of the experiments which, in 1930, resulted in 17-centimeter-wave transmission across the English Channel and of the developments for the Lympne-St. Inglevert microwave radiotelephone link which was inaugurated commercially in 1934. He was assistant director of research in 1945 when he was transferred to Federal Telecommunication Laboratories in New York as a consultant.

He has published extensively on high-frequency oscillators, wave guides, and general electromagnetic theory, and has taught field theory and applications of ultra high frequencies at the Ecole Supérieure d'Electricité.

Mr. Clavier is president of the section of the Société des Radioélectriciens dealing with hyperfrequencies. He is a Fellow of the Institute of Radio Engineers.

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J. H. Fremlin

J. H. Fremlin was born on March 4, 1913 at Kingsbury, Middlesex. He was educated at Berkhamsted School and Trinity College, Cambridge, where he obtained first-class honors in natural sciences in 1933. After working for three years in the Cavendish Laboratory under Lord Rutherford on discharges in gases and low-voltage neutron production, he was granted his Ph.D. in 1939. He entered the valve laboratory of Standard Telephones & Cables, Ltd. in 1937, where he did research work on triodes and later on velocity-modulated centimeter-wave oscillators in connection with the radar research program of the British Government. After the collapse of Japan, Dr. Fremlin was appointed Nuffield Research Fellow at the University of Birmingham, England. He is an Associate Member of the Institution of Electrical Engineers and a Fellow of the Institute of Physics.

William J. Gillule

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William J. Gillule was born on June 27, 1902. After graduating from the Marconi Institute in 1920, he served as a marine radio operator until 1926.

He joined the Federal Telegraph Company in 1926 to install and maintain radiotelegraph and direction-finder equipment. During 1928 and 1929, Mr. Gillule was transferred to Kolster Brandes Laboratories for work on the design of loudspeakers and associated sound equipment.

He was then transferred to the Marine Division of Mackay Radio and Telegraph Company and in 1934 was made chief inspector. During the following eleven years, Mr. Gillule collaborated in the design of the types 102, 105, and 106 direction finders. He directed the installation of communication and direction-finder equipment on several thousand liberty and victory ships, and tankers during the war. He was promoted to Production Manager of the Marine Division in 1945.

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R. N. Hall was born at Wigan, Manchester, in 1914 and was educated at Merchant Taylors School and Regent Street Polytechnic, London.

He joined the valve division of Standard Telephones & Cables Ltd., on production in 1932, and moved with it to the Sidcup location in 1933. In 1938 he joined the valve laboratories at Woolwich and subsequently moved to Eltham and Ilminster where he was engaged in developing centimetric valves in conjunction with the research establishments of His Majesty's Government.

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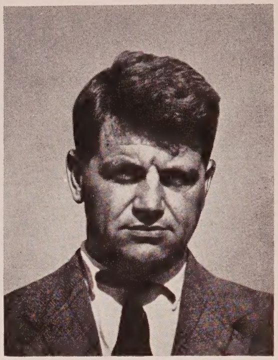
A. H. Hooke was born in Birmingham in 1917, and was educated at the Technical College, Brighton, Sussex. He joined the valve division of Standard Telephones & Cables, Ltd., early in 1938, and worked at both the Eltham and Woolwich locations prior to the move to Ilminster, Somerset.

Though an electrical engineer by profession, for which he passed the final City and Guilds electrical engineering and final City and Guilds radio communications examinations, Mr. Hooke also holds the National Diploma of the Institution of Mechanical Engineers. He has been chiefly engaged on valve measurements of all kinds, covering among other subjects ultra-high-frequency noise, power output in the centimeter-wave band, and cathode-ray-tube characteristics.



R. N. Hall

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R. F. Lewis was born in Watertown, New York, on January 5, 1908. He received a B.S. degree from Virginia Military Institute in 1930. He attended Wharton School of the University of Pennsylvania from 1931 to 1932.

A. H. Hooke

R. F. Lewis

He joined the RCA Manufacturing Company staff in 1933 and was transferred to the research division in 1935 where he was engaged in antenna studies until 1939. He then went to the Columbia Broadcasting System to work on television problems connected with antennas and radio-frequency filters. From 1942 to 1945, as a member of the Harvard Radio Research Laboratories, he was in charge of the antenna activities of the American-British Laboratories in England and on the continent.

He became a member of the engineering staff of Federal Telephone and Radio Corporation in 1945.

Mr. Lewis is an Associate Member of the Institute of Radio Engineers.

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F. J. Mann

F. J. Mann was born in Hoboken, New Jersey, on April 27, 1906. He received the B.S. degree from Middlebury College in 1933. Post-graduate studies were in communications and journalism, the latter at Columbia University School of Journalism.

From 1925 to 1927, he was on the staff of *Radio Broadcast* magazine. In 1928, he was appointed radio editor of *Science and Invention*. He served as a marine radio operator before completing college and after his post-graduate work he became a radio operator at broadcast station WCAX in Burlington, Vermont. From 1934 to 1940, he was engaged in newspaper work and free-lance writing. Mr. Mann then became assistant to the director of technical publications of Bendix Radio.

Joining the *Electrical Communication* staff as a writer in 1942, he is now managing editor of the publication and assistant editor of the handbook, *Reference Data for Radio Engineers*.

Mr. Mann is a member of the honorary journalism fraternity, Pi Delta Epsilon, and of the Institute of Radio Engineers.

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Philip A. Shatford was born in January, 1917, at Dartford, Kent. He attended the Paddington Technical Institute as a day student from 1931 to 1934.

He joined Standard Telephones & Cables, Ltd. at Ilminster, Somerset, in August 1941 and worked in the Valve Division on the development of velocity-modulated valves for centimeter wavelengths, being concerned mainly with continuously evacuated apparatus.

Mr. Shatford has recently been attached to the Standard Telecommunication Laboratories at Enfield in connection with a program of development in continuously pumped devices.

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P. A. Shatford

Norman H. Young was born in Philadelphia, Pennsylvania, in 1913. He received the B.S. degree in electrical engineering from Pennsylvania State College in 1934 and the M.S. degree in 1935.

From 1935 to 1942, he was engaged in television engineering for the Philco Corporation and had charge of the transmitter of television station WPTZ.

In 1942 he became a department head in Federal Telecommunication Laboratories and during the war was largely concerned with the application of pulse-time modulation to military communication equipment. At the termination of the war, he was responsible for the engineering of the color-television transmitter for the Columbia Broadcasting System. He has done additional work on receivers and studio equipment for color television.

Mr. Young is a member of Eta Kappa Nu and a senior member of the Institute of Radio Engineers.

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Norman H. Young

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

Associate Manufacturing and Sales Companies

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International Standard Electric Corporation, New York, New York

Federal Telephone and Radio Corporation, Newark, New Jersey

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International Marine Radio Company Limited, Liverpool, England

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Standard Electric Aktieselskab, Copenhagen, Denmark

Compagnie Générale de Constructions Téléphoniques, Paris, France

Le Matériel Téléphonique, Paris, France

Les Téléimprimeurs, Paris, France

Lignes Télégraphiques et Téléphoniques, Paris, France

Ferdinand Schuchhardt Berliner Fernsprech- und Telegraphenwerk Aktiengesellschaft, Berlin, Germany

Lorenz, C., A.G. and Subsidiaries, Berlin, Germany

Mix & Genest Aktiengesellschaft and Subsidiaries, Berlin, Germany

Süddeutsche Apparatefabrik Gesellschaft M.B.H., Nuremberg, Germany

Telephonfabrik Berliner A.G. and Subsidiaries, Berlin, Germany

Nederlandsche Standard Electric Maatschappij N.V., Hague, Holland

Dial Telefonkereskedelmi Részvény Társaság, Budapest, Hungary

Standard Villamossági Részvény Társaság, Budapest, Hungary

Telefongyár R.T., Budapest, Hungary

Fabbrica Apparecchiature per Comunicazioni Elettriche, Milan, Italy

Societa Finanziaria Elettrotelefonica, Milan, Italy

Societa Italiana Reti Telefoniche Interurbane, Milan, Italy

Nippon Electric Company, Limited, Tokyo, Japan

Sumitomo Electric Industries, Limited, Osaka, Japan

Standard Telefon- og Kabelfabrik A/S, Oslo, Norway

Standard Electric Company w. Polsce Sp. z.O.O., Warsaw, Poland

Standard Electrica, Lisbon, Portugal

Standard Fabrica de Telefoane si Radio S.A., Bucharest, Rumania

Compañía Radio Aerea Maritima Española, Madrid, Spain

Standard Eléctrica, S.A., Madrid, Spain

Aktiebolaget Standard Radiofabrik, Stockholm, Sweden

Standard Telephone et Radio S.A., Zurich, Switzerland

Jugoslavensko Standard Electric Company Akcionarno Drustvo, Belgrade, Yugoslavia

Teleoptik A.D., Belgrade, Yugoslavia

Telephone Operating Systems

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Compañía Telegráfico-Telefónica Comercial, Buenos Aires, Argentina

Compañía Telegráfico-Telefónica Del Plata, Buenos Aires, Argentina

Companhia Telefonica Paranaense S.A., Curitiba, Brazil

Companhia Telefonica Rio Grandense, Porto Alegre, Brazil

Compañía de Teléfonos de Chile, Santiago, Chile

Cuban Telephone Company, Havana, Cuba

Mexican Telephone and Telegraph Company, Mexico City, Mexico

Compañía Peruana de Teléfonos Limitada, Lima, Peru

Porto Rico Telephone Company, San Juan, Puerto Rico

Shanghai Telephone Company, Federal, Inc., U.S.A., Shanghai, China

Radiotelephone and Radiotelegraph Operating Companies

Compañía Internacional de Radio, Buenos Aires, Argentina

Compañía Internacional de Radio Boliviana, La Paz, Bolivia

Companhia Radio Internacional do Brasil, Rio de Janeiro, Brazil

Compañía Internacional de Radio, S.A., Santiago, Chile

Radio Corporation of Cuba, Havana, Cuba

Radio Corporation of Porto Rico, San Juan, Puerto Rico†

† Radiotelephone and Radio Broadcasting services.

Cable and Radio Telegraph Operating Companies

(Controlled by American Cable & Radio Corporation)

The Commercial Cable Company, New York, New York[1]

Mackay Radio and Telegraph Company, New York, New York[2]

All America Cables and Radio, Inc., New York, New York[3]

The Cuban All America Cables, Incorporated, Havana, Cuba[1]

Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[4]

[1] Cable service. [2] International and Marine Radiotelegraph services. [3] Cable and Radiotelegraph services. [4] Radiotelegraph service.

Laboratories

International Telecommunication Laboratories, Inc., New York, New York

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